Patent Application

of Steve Shattil

for

MULTIPLE INPUT, MULTIPLE OUTPUT CARRIER INTERFEROMETRY ARCHITECTURE

RELATED APPLICATIONS

This application is a Continuation In Part of U.S. Pat. Appl. 09/022,950 filed on February 12, 1998, which is now U.S. Pat. No. 5,955,992.

BACKGROUND OF THE INVENTION

I. Field of the Invention

The present invention relates to a multiple-input, multiple-output (MIMO) communication system that uses a novel multicarrier spread-spectrum protocol.

II. Description of the Related Art

Multi-Carrier Modulation (MCM) is the principle of transmitting data by dividing the stream into parallel low bit-rate streams, then using each substream to modulate a carrier.

Orthogonal frequency division multiplexing (OFDM) is an MCM method that transmits different data streams over multiple *n* equally spaced carrier frequencies. Since the carrier-frequency channels are narrowband, they experience flat fading, which simplifies equalization. High spectral efficiency is obtained because the carrier channels overlap in the frequency domain. Fourier-transform processing is used for modulation and demodulation, and information is carried in the phase shift applied to each carrier. The value of *n* is typically a power of two in order to facilitate computation of a sampled version of the signal with a fast Fourier transform (FFT). OFDM systems may use coding to mitigate bit errors on subcarriers that are in deep fades. This reduces the bandwidth efficiency of OFDM as the number of subcarriers becomes larger than the number of transmitted symbols.

Peled and Ruiz ("Frequency domain data transmission using reduced computational complexity algorithms," Proc. IEEE Int. Conf. Acoust., Speech, Signal Processing, p 964-967, Denver, CO, 1980) introduced a cyclic prefix to OFDM. A guard space between OFDM symbols is filled with a cyclic extension of the symbol to simulate the performance of a cyclic convolution.

In "Performance analysis of coded OFDM on fading channels with non-ideal interleaving and channel knowledge" (which is hereby incorporated by reference) Sandell et. al. mentions the use of a cyclic prefix to preserve the orthogonality of the OFDM subchannels and mitigate intersymbol interference between consecutive OFDM symbols.

In "A Multicarrier Primer," Cioffi describes a multicarrier modulation scheme in which the number of data bits allocated to different subchannels may be varied relative to the attenuation and noise on each channel. Cioffi also describes improvements to Discreet Multitone Modulation.

Another Stanford University faculty member, J. Campello, describes multicarrier coding techniques in "Optimal Discrete Bit Loading for Multicarrier Modulation Systems", Proceedings of the 1995 IEEE International Symposium on Information Theory.

In U.S. Pat. No. 5,471,647, Gerlach and Paulraj describe a transmission method using an antenna array and probing signals together with feedback from receivers to the array. This probing-feedback method allows the transmitter to estimate the instantaneous channel vector, even when there is time-varying multipath in the propagation medium.

In U.S. Pat. No. 5,634,199, Gerlach et. al. describe adaptive transmitting antenna methods based on an autocorrelation matrix associated with each receiver instead of using the receiver's instantaneous channel vector. The autocorrelation matrix, which describes the time-averaged behavior of the channel vector, requires far less feedback to track. In the '199 patent, Gerlach mentions an "open loop" approach explored by G. Raleigh et. al. ("A Blind Adaptive Transmit Antenna Algorithm for Wireless Communication," International Communications Conference, 1995) as less robust to changes in the propagation medium than feedback methods.

In U.S. Pat. No. 5,671,168, Liu describes one of many well-known array-processing techniques that separate received signals into frequency components and apply adaptive beam-forming weights to those components.

In U.S. Pat. No. 5,528,581, De Bot describes array processing in which each received signal from an array element is separated into subchannel frequencies. Each of the subchannel frequencies is combined in a separate processor to enable spatial-diversity benefits. OFDM is described as one particular transmission protocol that benefits from separate spatial processing within each subchannel frequency bin.

In PCT Appl. No. PCT/US94/08247, Shattil describes interference cancellation techniques that compensate for propagation path differences and differences in receiver responses to interfering multi-frequency signals. Adaptive weights may be frequency-dependent to optimize cancellation of signals having multiple frequencies.

In U.S. Pat. No. 6,008,760, Shattil describes interference cancellation methods for antenna arrays in which weights are determined from measured ratios of interference.

In U.S. Pat. No. 6,144,711, Raleigh describes an MIMO system and method that is no more novel than the sum of already well-known techniques borrowed from others. Raleigh re-introduces the use of well-known multicarrier array processing (e.g., De Bot, Liu, Shattil) for OFDM. Raleigh reiterates feedback and autocorrelation matrix techniques (e.g., Gerlach, Paulraj). Raleigh reacquaints the reader with bit loading and

coding for fading channels (e.g., Cioffi, Campello) and repeats the use of cyclic prefixes (e.g., Sandell, Peled and Ruiz). Raleigh provides in-depth discussions on interleaving and error-check coding, which are also well known in the same field.

In practice, Raleigh's "invention" (marketed by Cisco Systems as "V-OFDM") achieves substantially inferior performance benchmarks compared to Lucent Technologies' "Bell Labs Layered Space Time" (BLAST) system, which is based on related, but novel technologies. None of the prior-art references describe an MIMO system that benefits from the combination of spatial processing and an Interferometry-based multicarrier protocol.

Other MCM techniques have been developed. For example, Multicarrier CDMA (MC-CDMA) is a combination of OFDM and CDMA. Unlike conventional direct-sequence CDMA (DS-CDMA), MC-CDMA employed with a spreading factor of N can accommodate N users with good BER without requiring interference cancellation. MC-CDMA avoids OFDM encoding by using an NxN matrix operation. In multi-tone CDMA (MT-CDMA), the available spectrum is divided into a number of equiwidth frequency bands that are used to transmit a narrowband direct-sequence waveform.

U.S. Pat. Nos. 5,519,692 and 5,563,906 describe geometric harmonic modulation (GHM) in which preamble and traffic waveforms are created from multiple carrier frequencies (tones). The waveforms comprise tones incorporating a binary phase code where signal phases are 0 or $-\pi/2$. The binary phase offsets, which are applied to the tones, provide the spreading codes. Orthogonality of GHM signals is realized upon correlation with a reference signal at a receiver. A preamble carrier waveform is constructed by summing the tones. Therefore, the preamble signals are similar to MC-CDMA signals.

Each receiver monitors the preamble signals for its phase code and then despreads and decodes the appended traffic waveforms. The traffic waveforms are products of the tones. The receiver generates a reference waveform from a product of tones having phase offsets that correspond to the receiver's phase code. The reference waveform is correlated with the received signals to produce a correlation result that is integrated over the data-bit duration and over all tones.

GHM uses binary phase offsets instead of differential phase offsets. Thus, GHM does not provide carriers with phase relationships that enable the superposition of the carriers to have narrow time-domain signatures or orthogonality. Consequently, received GHM signals require processing by a correlator, whereas signals that are orthogonal in time can be processed using simpler signal-processing techniques, such as time sampling and weight-and-sum. Furthermore, GHM does not achieve the capacity and signal-quality benefits enabled by interferometry-orthogonal signals.

U.S. Pat. No. 4,628,517 describes a radio system that modulates an information signal onto multiple carrier frequencies. Received carriers are each converted to the same intermediate frequency using a bank of conversion oscillators. The received signals are then summed to achieve the benefits of frequency diversity. In this case, frequency diversity is achieved at the expense of reduced bandwidth efficiency. The process of converting the received signals to the same frequency does not allow orthogonality between multiple information signals modulated on the same carriers.

Many conventional diversity-combining techniques combine several received signals to improve signal-to-noise ratios prior to demodulation. However, in U.S. Pat. No. 5,548,819, Robb describes a receiver that correlates copies of demodulated information signals. The '819 receiver uses correlation between multiple information signals, each containing a copy of the message of interest. In the '819 transmitter, each information signal is modulated onto a plurality of carriers with the constraint that no more than half of the carriers be shared with any other information signal.

The methods described in the '819 patent are effective only under the assumption of uncorrelated interference. Furthermore, the system described is quasi-orthogonal at best. Multiple-access interference increases as more users are forced to share carriers. The '819 patent fails to recognize the use of carrier phase spaces to orthogonalize signals sharing the same carriers, which can be achieved without the carrier-assignment constraints described in the '819 patent.

Each conventional multicarrier transmission protocol presents different benefits and disadvantages. Benefits can be increased by merging different protocols, but only to a limited degree. However, Carrier Interferometry (CI) can achieve all of the performance benefits of each of transmission protocol. CI can be used as an underlying architecture for

single-carrier, multicarrier, and wideband transmission protocols. Thus, it is preferable that an MIMO method be developed from the CI multicarrier underlying architecture that enables any transmission protocol to realize the benefits of MIMO.

SUMMARY OF THE INVENTION

Disregarding Cisco Systems' claims to have "invented" the only MIMO method, this invention describes a class of MIMO techniques and systems that includes dozens of unique embodiments.

Carrier Interferometry Multiple Access (CIMA) is a class of multiple-access techniques that use sets of phase shifts (i.e., phase spaces) to overlay and separate data streams that are redundantly modulated onto the same sets of carrier signals. Multiple carriers are redundantly modulated with data streams that are orthogonalized from each other by virtue of the different sets of phase spaces encoding each data stream. Interference between the carriers provides the means to orthogonalize the data streams, whether the carriers are combined or processed separately.

Modulated carriers representing a particular data stream may be combined to produce a superposition signal that is orthogonal or quasi-orthogonal in time relative to superposition signals representing other data streams. Each of the modulated carriers corresponding to a particular data stream may be processed separately. Carriers modulated with other data streams may contribute some interference to the processing. The processing may be adapted to provide destructive combining (cancellation) of the interference, or the interference may be mitigated via decision processing.

Carrier Interferometry (CI) describes the applications of a redundantly modulated multicarrier architecture to conventional transmission protocols, such as TDMA, CDMA, and OFDM. CI uses a multicarrier architecture and a selection of individual carrier characteristics (such as phase, frequency, and amplitude) to generate desired time-domain characteristics from a superposition of the carriers. CI provides frequency diversity, mitigates interference, increases capacity, and achieves additional benefits described in the preferred embodiments and in several of the documents included by reference.

Since CI can preserve the time-domain characteristics of conventional protocols, both single-carrier and multicarrier processing may be performed on CI-based signals.

Thus, CI enables a seamless integration of high-performance multicarrier processing in communication networks that use conventional transmission protocols.

A CI procedure includes signal-processing operations performed at either or both transmit and receive sides of a communication channel. In one embodiment, a CI procedure ensures that each of a plurality N' of input symbols is modulated (or otherwise impressed) onto a plurality N of frequency bins such that interference between the bins enables separation of each of the N' input symbols. For example, interference between the frequency bins causes constructive interference of a first input symbol and destructive interference of the second through the N'^{th} input symbols within a predetermined time interval.

CI is distinguishable from conventional MCM techniques because it achieves both frequency diversity and bandwidth efficiency. CI redundantly transmits each data bit over multiple carriers that may or may not be equally spaced in frequency. Data channels are orthogonalized by a set of phase shifts applied to the carriers. Information is generally modulated onto the carriers via phase-shift or pulse-amplitude modulation.

Sensitivities to phase jitter and receiver frequency offsets are recognized as the major disadvantage of MCM. However, CI is substantially insensitive to such sources of interchannel interference.

CI enables direct up-conversion (for transmission) and direct down-conversion (for reception), and thus, eliminates the need for RF and IF components in transceivers. Ultra-wideband implementations of CI are possible using slow, parallel-processing techniques.

CI can be implemented as an underlying signal architecture for conventional protocols (such as CDMA, TDMA, and OFDM). The benefits of CI implementations for other protocols include unprecedented capacity increases and signal-quality improvements, as described in "MMSE Frequency Combining for CI/DS-CDMA," (IEEE RAWcon, Denver, CO, 2000), "Introduction of Carrier Interference to Spread Spectrum Multiple Access" (IEEE Emerging Technologies Symposium, Richardson TX, 1999), "Exploiting Frequency Diversity through Carrier Interferometry" (Wireless 2000), "Throughput Enhancement in TDMA through Carrier Interference Pulse Shaping (VTC 2000), "Wireless Communication System Architecture and Physical Layer Design for

Airport Surface Management" (VTC 2000), and "Array Control Systems For Multicarrier Protocols Using A Frequency-Shifted Feedback Cavity" (IEEE RAWcon, Denver, CO, 1999), which are incorporated by reference. CI reduces the cost and complexity of conventional communication systems and provides a simple overlay to existing systems that allows a seamless transition from conventional communication protocols to CI-based protocols.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1A illustrates CI carriers having phase fronts that are aligned at a specific time.

FIG. 1B illustrates an in-phase superposition of CI carriers.

FIG. 2A shows a plurality of carrier signals defined by a range of carrier frequencies and sets of phase offsets, such as phase offsets resulting in a particular time-domain characteristic of a superposition of the carriers.

FIG. 2B shows a plurality of carrier signals defined by a range of carrier frequencies, sets of phase offsets, and sets of carrier amplitudes and a superposition of the carriers. Either or both the superposition signal and the set of carriers may be processed for transmission and reception.

FIG. 3 shows a plurality N of carrier signals that are redundantly modulated with a stream of data symbols $s_k(t)$ associated with a k^{th} channel.

FIG. 4A illustrates the use of different phases and time intervals applied to one set of carrier frequencies to distinguish a k^{th} channel.

FIG. 4B represents a plurality of data symbols $s_1(t)$, $s_2(t)$, and $s_3(t)$ modulated onto the same carrier frequencies within the same time interval.

FIG. 5 illustrates a plurality of time-offset carrier signals having different initial time offsets t^o_{kn}.

FIG. 6A shows a plurality of carriers corresponding to a plurality M of subchannels each having a plurality N of carriers.

FIG. 6B illustrates a superposition of three subchannels that occupy the same time-domain space.

FIG. 7 illustrates a time-offset code used with carrier-defined sub-channels.

FIG. 8A illustrates a method of transmitting CI signals.

FIG. 8B shows steps in a CI transmission method.

FIG. 8C illustrates a method of generating CI signals.

FIG. 9A illustrates a method of receiving and processing CI signals.

FIG. 9B illustrates a method of receiving CI signals.

FIG. 10 illustrates a functional diagram of a system that generates CI signals.

FIG. 11 is a diagram of a CI transmission system.

FIG. 12A shows an embodiment of CI transmission system.

FIG. 12B shows an embodiment of CI transmission system.

FIG. 12C illustrates a CI-CDMA transmitter with respect to how it functions.

FIG. 13A shows a basic CI receiver for a k^{th} user.

FIG. 13B shows a basic CI receiver for a k^{th} user.

FIG. 14A shows an embodiment of CI receiver configured to operate with a DS-CDMA protocol.

FIG. 14B shows an embodiment of one of a chip receiver shown in FIG. 14A.

FIG. 15 shows a spatial-interferometry receiver that includes a beam-forming stage and an interference-cancellation stage.

FIG. 16 is a diagram that shows a pair of orthogonal linear-polarization directions.

FIG. 17 shows a polar plot of intensity versus polarization offset φ relative to a receiver's polarization. The polar plot is regarded as a beam pattern in polarization space for a polarized receiver.

FIG. 18A shows a polar plot of two polarization beam patterns wherein each beam pattern corresponds to one of two polarized transceivers.

FIG. 18B shows a polar plot of a polarization beam pattern resulting from the weighted sum of the beam patterns shown in FIG. 18A.

FIG. 19A shows a pair of linear-polarized directions having a relative polarization offset of φ.

FIG. 19B shows a pair of circular (or elliptical) polarized directions having identical angular velocities ω and a polarization offset of φ.

FIG. 19C shows a pair of circular (or elliptical) polarized directions having different angular velocities ω and ω'.

FIG. 20A shows a plot of a polarization offset $\varphi(t)$ over a period $T = 1/\Delta\omega$ resulting from a pair of circular polarized directions having different angular velocities ω and ω'.

FIG. 20B shows a pair of polarization offsets $\varphi(t)$ and $\varphi'(t)$ over a period $T = 1/\Delta\omega$ resulting from two pairs of circular polarized directions having different initial polarization offsets φ_o and φ'_o.

FIG. 20C shows a plurality N of polarization offsets $\varphi(f,t)$ having different frequencies f_i. In this case the frequencies f_i are harmonics. However, other frequency relationships may be selected.

FIG. 21A shows a pair of gain distributions $g_1(f)$ and $g_2(f)$ in frequency space for a pair of receivers having different frequency responses.

FIG. 21B shows a pair of gain distributions $g_1(\chi)$ and $g_2(\chi)$ for a non-spatial diversity parameter space (χ-space) for a pair of receivers having different responses in the χ diversity-parameter space.

FIG. 22A shows two sinusoidal signals having a similar frequency f_1 and phase ϕ_1. A correlation of the signals yields a normalized value of one.

FIG. 22B shows two sinusoidal signals having orthogonal frequencies f_1 and f_2. If the signals are correlated over an interval of $T_s = 1/(f_1-f_2)$, the correlation yields a value of zero.

FIG. 22C shows two sinusoidal signals having a similar frequency f_1 but different phases ϕ_1 and $\phi_{n\neq1}$. If these signals are correlated over an interval of $T_s = 1/(f_1-f_2)$, the correlation yields a normalized value between 1 and −1.

FIG. 22D shows two sinusoidal signals having a similar frequency f_1 but different phases ϕ_1 and $\phi_1 + \pi/2$. The $\pi/2$ phase difference is a special case known as in phase and quadrature. The correlation of these signals yields a value of zero.

FIG. 23 shows a functional diagram of a phase-space beam-forming system.

FIG. 24 shows a superposition of 11 phase-locked carrier frequencies modulated with an information symbol.

FIG. 25A through FIG. 25K show correlations between two orthogonal phase spaces $\Phi_n(f_i, \phi_j)$ for eleven groups of carrier frequencies f_i. Each figure shows two carriers

having the same frequency. The carrier frequency shown in any one figure is related to carrier frequencies in any other figures by an integer-multiple shift of a separation frequency f_s. In this case, the phase difference between the carriers shown in each figure is related to a phase shift associated with a translation in time. The amount of the translation in time is similar for each pair of carrier frequencies. However, because the frequencies are different for each pair of carriers, the same translation in time can result in different relative phases between each pair of carriers.

FIG. 26 illustrates a transmitter in accordance with one embodiment of the invention.

FIG. 27 illustrates a CI receiver according to one embodiment of the invention.

FIG. 28A illustrates a wireless communication system of the present invention wherein a CI/spatial processing transmission system is coupled to a transmitter array and a CI/spatial processing receiver system is coupled to a receiver array.

FIG. 28B shows a transmit CI/spatial processing system coupled to an antenna array and a receiver CI/spatial processing system coupled to an antenna array wherein each system includes a beam-forming system.

FIG. 28C shows a transmit CI/spatial processing system coupled to an antenna array and a receiver CI/spatial processing system coupled to an antenna array wherein the arrays each having a plurality of polarizations.

FIG. 28D shows a generalized communication system of the present invention wherein a transmitter generates a set of CI carrier signals modulated by a plurality N' of data-symbol streams $s_{n'}(t)$. An n^{th} carrier is distinguished by at least one diversity-parameter value Λ_n.

FIG. 28E shows a plurality of transmission systems separated with respect to at least one diversity parameter.

FIG. 29A shows a plurality of waveforms that each has a similar frequency and a different phase.

FIG. 29B shows a correlation of a reference signal with a same-frequency received signal. The correlation is shown over a phase-offset interval of 2π.

FIG. 30 shows a communication system according to one embodiment of the invention.

FIG. 31A shows a plurality of spatially separated transmitters and a receiver having an array of receiver elements and a beam-forming system.

FIG. 31B shows a communication system in which a transmitter includes a plurality of transmitter elements and a receiver includes an antenna array and a beam-forming system.

FIG. 32 shows a plurality N of carrier frequencies wherein each carrier frequency f_n (n=1,...,N) has a plurality of phased subcarriers.

FIG. 33A shows a plurality N of incrementally spaced-in-frequency carrier signals that may be used as an underlying architecture for transmission protocols. Users and/or data channels having symbols that are redundantly modulated on all N carriers can be provided with N orthogonal phase spaces.

FIG. 33B shows an arrangement of N incrementally spaced-in-frequency carrier signals divided into M sets of carrier signals. Each set may include one or more users and/or data channels. Each set may or may not include a similar number of carriers. The carriers in each set may be incrementally spaced or non-incrementally spaced.

FIG. 34A shows a single carrier frequency f_n having multiple quasi-orthogonal phases $\phi_{(mn)k}$ that can be used in a quasi-orthogonal phase-space division multiplexing technique.

FIG. 34B shows a plurality of carrier frequencies wherein each frequency includes multiple quasi-orthogonal phases.

FIG. 35A shows a serially transmitted high-bandwidth signal sampled over a sample-time interval.

FIG. 35B shows a frequency-domain plot of the signal shown in FIG. 35A.

FIG. 35C shows a decomposition of the signal shown in FIG. 35A into N carrier-frequency components.

FIG. 35D shows a frequency-domain plot of the signal shown in FIG. 35C.

FIG. 35E shows a decomposition of the signal shown in FIG. 35A into N' carrier components where $N' > N$.

FIG. 35F shows a frequency-domain plot of the signal shown in FIG. 35E.

FIG. 36 illustrates a CI process that may be used to process various types of wideband signals by decomposing the signal into discreet components and separating

transmitted data symbols embodied in the values of the components. Any type of multi-user detection may be used to separate the transmitted data symbols. Constellation methods may be used with, or in place of, other multi-user detection methods.

FIG. 37A illustrates a CI-TDMA signal constructed from time-shifted and scaled versions of a basic CI pulse.

FIG. 37B shows basic CI pulse, which may be used to generate time-domain signals that correspond to conventional protocols, such as TDMA.

FIG. 37C illustrates how multiple closely spaced or overlapping pulses may be combined to generate longer-duration signals.

FIG. 38 shows a CI-OFDM signal architecture for a plurality M of users wherein each user is provided with a unique set of carriers.

FIG. 39 shows a CI MC-CDMA architecture in which a k^{th} user having a k^{th} phase space ϕ_{kn} is provided with a spreading sequence c_{kn} assigned to each of a plurality N of carriers. The spreading sequence c_{kn} may include binary or differential values.

FIG. 40A shows a time-domain representation of a plurality K of DS-CDMA codes generated from a CI architecture. Each chip of each CI DS-CDMA code is generated from a CI superposition of carriers. Each chip is provided with a binary code value corresponding to a DS-CDMA spreading code.

FIG. 40B illustrates chip overlap, which can increase bandwidth efficiency and or provide additional redundancy. Chip overlap introduces a quasi-orthogonal condition. In the example shown, chip overlap doubles the code length without increasing bandwidth.

FIG. 40C shows one of many possible carrier architectures for a CI DS-CDMA system.

FIG. 41 shows a CI architecture for a Frequency-Hopping Spread Spectrum transmission protocol.

FIG. 42 shows $K{\cdot}M$ symbols $s_{km}(t)$ modulated onto same-frequency carrier signals. A plurality K of channels or users each modulate a plurality M of symbols onto the carriers relative to a plurality M of phases ϕ_{kmn}.

FIG. 43A shows K sets of M symbols $s_{km}(t)$ ($k = 1,\ldots,K$ and $m = 1,\ldots,M$) coupled into a plurality K of transmitter spatial weight (TSW) systems in a transmitter of the invention.

FIG. 43B shows a transmitter having TSW's that weights sets of symbols before the symbols are modulated onto the carriers.

FIG. 43C shows a transmitter of the invention in which information-modulated carriers are weighted.

FIG. 44A shows a CI-spatial receiver of the invention.

FIG. 44B shows an embodiment of a multi-user detector that may be used in a CI-spatial receiver of the invention.

FIG. 44C shows a matched-filter embodiment of a decision/detector system.

FIG. 44D shows an embodiment of a multi-user detector that may be used in a receiver that processes redundantly modulated carriers.

FIG. 44E shows an embodiment of a section of a multi-user detector that supports phase-encoded, non-redundantly modulated carrier frequencies.

FIG. 45 illustrates functional components of a field-programmable gate array (FPGA) chip that was constructed as a signal-processing core for a CI transceiver.

FIG. 46A illustrates an FPGA-generated time-domain CI pulse, which is formed by a superposition of ten unweighted carrier frequencies that are spaced about 100 kHz apart.

FIG. 46B is a frequency-domain plot of a CI pulse shown in FIG. 46A.

FIG. 47A illustrates a simplified DS-CDMA signal that was formed using a 32-chip Hadamard-Walsh code with 32 carriers.

FIG. 47B shows relative magnitudes of frequency components that comprise the signal shown in FIG. 47A.

FIG. 48 illustrates a frequency-domain plot of a transmitted 32-chip Hadamard-Walsh signal in a multipath environment.

FIG. 49A shows a time-domain plot of a 64-chip Hadamard-Walsh code with overlapping pulses generated from 32 carriers.

FIG. 49B shows a frequency-domain plot of the time-domain signal shown in FIG. 49A.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

1. Definitions

Various terms used in the descriptions of CI methods and systems are generally described in this section. The description in this section is provided for illustrative purposes only, and is not limiting. The meaning of these terms will be apparent to persons skilled in the relevant art(s) based on the entirety of the teachings provided herein. These terms may be discussed throughout the specification and the cited references with additional detail.

The term carrier signal or carrier, when used herein, refers to at least one electromagnetic wave having at least one characteristic that may be varied by modulation. Other wave phenomena, such as acoustic waves, may be used as carriers. Carrier signals may include any type of periodic signal. Carrier signals may include sinusoids, square waves, triangle waves, wavelets, and/or arbitrary waveforms. A carrier signal is capable of carrying information via modulation. A carrier signal may be modulated or unmodulated. Multicarrier signals may include multi-frequency signals, time-domain (discreet-time) signals, and/or any other set of electromagnetic signals having different values of at least one diversity parameter.

A coder or encoder, as used herein, includes any algorithm, method, component, system, and/or adapted use of natural phenomena that performs coding of an information-bearing signal.

Coding or encoding a signal, as used herein, can involve distributing symbols among a plurality of diversity-parameter values. A code refers to any type of digital or analog code applied to at least one information signal and/or carrier signal. For example, an encoder may encode symbols across available time intervals, carrier frequencies, phase spaces, spatial subchannels, polarizations, circular polarization rotation rates, and/or spatial directions.

Coding may be performed based on estimation of at least one communication channel, responsiveness to known training symbols, and/or optimization of at least one measurement (such as signal to noise-plus-interference). Coding may include spatial processing and/or non-spatial diversity parameter-space processing. Coding may include

virtual addressing. Coding may include CI processing, such as applying weights to carrier signals to provide for encryption, error correction, error checking, redundancy, diversity enhancement, spreading, windowing, multiple access, or the like. Coding may include any type and combination of convolutional encoding, interleaving, trellis encoding, Reed-Solomon encoding, Viterbi encoding, ARQ coding, Hadamard-Walsh coding, Maximal-length coding, Gold coding, turbo coding, block coding, error-correction coding, modulation, and/or adaptive encoding to modify bit and/or power loading relative to diversity-parameter values and/or diversity-parameter subspaces. Coding may include scaling information symbols with respect to a predetermined constellation of values. In some cases (such as in coherence multiplexing, frequency-diversity interferometry, spatial interferometry multiplexing, etc.), a communication channel can encode a transmitted signal. Multipath effects, dispersion, and various types of distortion can encode or enhance encoding of transmitted signals.

The term communication channel, as used herein, may be natural and/or man-made including, but not limited to, air, space, wire, cable, waveguide, microstrip, strip-line, optical fiber, liquid, etc. A communication channel may include any propagation medium and/or path between at least one transmitter and ant least one receiver.

The term coupler, when used herein, can include one or more of the following: an antenna, an optical coupler, a modem, connector, or any other device that can be used to interface with communications medium. A coupler may include any type of transducer including transducer arrays. A transducer includes transmitters, receivers, and/or any device or system that acts as both a transmitter and a receiver. A coupler may include one or more signal-processing devices including, but not limited to amplifiers, filters, up-converters, down-converters, modulation-protocol converters, coding systems, decoding systems, mixers, delay systems, polarizers, phase shifters, delay systems, beam-forming systems, interferometers, multi-user detectors, cancellation systems, and signal combiners.

The term coupling, when used herein with reference to coupling a signal into a communication channel, refers to any method of inserting an electromagnetic signal into a communication channel. Coupling may include one or more processing steps including,

but not limited to up-converting, down-converting, filtering, weighting, coding, amplifying, mixing, delaying, combining, and polarizing.

The term coupling, when used herein with reference to coupling a signal out of a communication channel, may include one or more processing steps including, but not limited to, demodulating each of the received carriers, weighting one or more of the received carriers, down-converting received signals, up-converting received signals, converting received signals to a different modulated protocol, canceling interference, filtering, amplifying, decoding, and analyzing the received carriers. Coupling may involve signal-processing methods after a superposition or combining process, such as, but not limited to, error detection, decoding, filtering, windowing, amplification, interference cancellation, and multi-user detection.

A decoder, as used herein, includes any algorithm, method, component, system, and/or adapted use of natural phenomena that performs decoding of a coded signal.

Decoding a signal, as used herein, typically involves providing an inverse relationship to a coding process in order to extract data symbols from an encoded data stream. Typically, decoding is performed by a decoder on a receive end of a communication channel. Decoding may include multi-user (multi-channel) detection to separate interfering signals. Decoding may include constellation methods to determine the values of interfering signals. In some cases (such as in coherence multiplexing, frequency-diversity interferometry, spatial interferometry multiplexing, etc.), a communication channel can be used to decode a transmitted signal. Multipath effects, dispersion, and various types of distortion can decode or enhance decoding of a transmitted signal. Decoding may include well-known reception techniques (such as maximum likelihood detection). Decoding may include feedback loops to enhance at least one measured value.

The term diversity-parameter, as used herein, defines a signal characteristic that enables a signal to be distinguished from another signal. Examples of diversity parameters include, but are not limited to amplitude, space, directionality, energy, power, polarization direction, circular/elliptical polarization rotation rate, mode, frequency, time, code, phase, coherence length, and phase space. Diversity parameters may include proportions of two or more diversity-parameter values. Diversity parameters may include

any combination of unique signal characteristics. Diversity parameters may include diversity-parameter subspaces, such as spatial subspaces. A common diversity parameter, as used herein, is a range of at least one diversity-parameter value into which electromagnetic signals may be mapped.

The term diversity parameter space, as used herein, describes a set or range of values of at least one diversity parameter. For example, a plurality of receivers adapted to have different responsiveness to different values of at least one diversity parameter can process received signals with respect to the diversity-parameter space(s) in which the receivers are adapted. A diversity-parameter space may include any combination of diversity-parameter spaces and subspaces.

The term electromagnetic signal, when used herein, refers to any signal(s) in the electromagnetic spectrum. The electromagnetic spectrum includes all frequencies greater than zero hertz. Electromagnetic waves generally include waves characterized by variations in electric and magnetic fields. Such waves may be propagated in a communication channel. An electromagnetic wave may refer to an electrical signal, movement of charges, or changes in an electrical potential.

The term information signal (or baseband signal), when used herein, is an electromagnetic signal that includes, but is not limited to, video baseband signals, voice baseband signals, computer baseband signals, etc. Baseband signals include analog baseband signals and digital baseband signals. An information signal may be a coded signal that is coded with one or more codes. An information signal may be an information-bearing signal. For example, an information-bearing signal may be an intermediate-frequency, effective-carrier, and/or subcarrier signal modulated with at least one information signal.

The term modulation, when used herein, refers to any of a variety of techniques for impressing information from one or more baseband signals onto one or more carrier signals. The resulting signals are referred to as modulated carrier signals. Modulation imparts changes to the carrier signal that represent information in a modulating baseband signal. The baseband signal may be coded. The changes can be in the form of changes to one or more diversity parameters that characterize the carrier signal. A carrier signal can be modulated with a plurality of modulation types. Modulation of the carrier signals can

be performed with any type of modulation including but not limited to: phase modulation, amplitude modulation, frequency modulation, time-offset modulation, polarization modulation, or any combinations thereof. A carrier signal may be modulated with a plurality of baseband signals, such as analog baseband signals, digital baseband signals, coded baseband signals, and combinations thereof. Modulation may include shift key, diversity parameter difference, and/or continuous modulation techniques.

A modulator performs modulation, as defined herein. A modulator may redundantly modulate an information signal onto a plurality of carrier signals to generate a redundantly modulated multicarrier signal. A modulator may redundantly modulate a plurality of information signals onto a plurality of carriers to generate a plurality of redundantly modulated multicarrier signals wherein each multicarrier signal is a plurality of carriers redundantly modulated with at least one information signal.

A modulator may modulate a plurality of information signals onto one carrier to generate a plurality of modulated carrier signals. In this case, modulation transforms the carrier signal into a plurality of distinguishable carrier signals when the carrier phase of each modulated carrier is provided with a predetermined phase relationship that facilitates separation of the information signals. Multi-user (or multi-channel) detection techniques can be used to separate distinguishable, yet interfering signals. The carrier also may be regarded as an overloaded carrier since each of the distinguishable carrier signals shares the same frequency. Overloading also can be accomplished by adjusting some other diversity parameter(s) (e.g., polarization, spatial characteristics, etc.) of the carrier signal. Overloading enables frequency reuse and thus, enhanced bandwidth efficiency. CI involves overloading redundantly modulated carriers. Carrier phase spaces enable orthogonality between information signals in overloaded carriers if the number of phase spaces is less than or equal to the number of carriers. Otherwise, the phase-space channels are quasi-orthogonal.

Receiver element or receiver, as used herein, refers to any device that is capable of being responsive to electromagnetic radiation. A receiver is typically adapted to be coupled to a communication channel. A receiver may include a plurality of receiver elements. Receiver elements are typically distinguished from each other by having different responsiveness relative to at least one diversity parameter. For example,

spatially separated receiver elements are capable of having different phase responses and/or magnitude responses to received signals due to the locations of the receiver elements. The receiver responses can vary with respect to differences in spatial characteristics of the received signals (as well as the propagation environment), such as directivity of the received signals, multipath effects, and location of signal sources. A plurality of receiver elements may share a common receiver structure, such as an antenna. However, each receiver element includes at least one processing element (such as a filter, a correlator, a beam-forming circuit, a detector, a sampler, etc.) that allows the receiver element to be distinguished from other receiver elements. Different signal-processing algorithms can be regarded as different receiver elements if they provide multiple responses to received signals.

Spatial processing, as used herein, includes any combination of techniques that mitigate interference between signals having overlapping and/or similar diversity-parameter values. Spatial processing may include spatial interferometry, beam forming, and/or array-processing techniques that may be adapted to one or more diversity-parameter spaces.

The term subcarrier, as used herein, refers to any type of periodic signal and/or code signal. A subcarrier may include more than one signal and more than one type of signal.

The term superposition signal, as used herein, refers to a signal produced by combining (e.g., summing) multiple component (e.g., carrier) signals. The carrier signals may be weighted with scalar and/or complex values before being combined to produce a superposition signal. Weighting may include multiplication by a scalar or complex value, phase shifting, delay, attenuation, amplification, gain adjustment, and/or any other method of changing the magnitude and/or phase of at least one signal. A superposition signal is typically a time-domain signal (i.e., a signal having some predetermined and/or desirable time-domain characteristic(s)).

The term transmission protocol, as used herein, describes any type of signaling protocol used for wireless and/or waveguide communications. A transmission protocol may include any type of spread spectrum. A transmission protocol may include one or

more multiple-access protocols, as well as protocols used for single-channel and broadcast communications.

Transmitter element or transmitter, as used herein, refers to any device that is capable of generating at least one information-bearing electromagnetic signal. A transmitter is typically capable of being coupled to a communication channel. A transmitter may include a plurality of transmitter elements. Transmitter elements are typically distinguished from each other functionally by their ability to provide transmitted signals having distinct values or characteristics of at least one diversity parameter. For example, transmitter elements may be spatially separated, thus providing transmitted signals with spatially dependent magnitudes and phases. Transmitter elements may be distinguished from each other physically by having separate components, although transmitter elements may share the same components and systems. For example, multiple beam-forming circuits may share the same antenna array. Each beam-forming circuit can be regarded as a transmitter element because it is capable of providing unique transmit characteristics to the transmitted signal. Different signal-processing algorithms can be regarded as separate transmitter elements if the algorithms provide multiple transmit signals with different characteristics of at least one diversity-parameter value.

2. Introduction to CIMA and CI Processes

The Description of the preferred embodiments assumes that the reader has a familiarity with CI. U.S. Pat. No. 5,955,992, U.S. Pat. No. 6,008,760, PCT Appl. No. PCT/US99/02838, PCT Appl. No. PCT/US00/18113, U.S. Pat. Appl. No. 09/347,182, U.S. Pat. Appl. No. 324,206, U.S. Pat. Appl. No.09/472,300, U.S. Pat. Appl. No.09/433,196, U.S. Pat. Appl. No. 09/393,431 U.S. Provisional Pat. Appl. No. 60/163,141, U.S. provisional Pat. Appl. "Method and Apparatus for Transmitting and Receiving Signals having a Carrier Interferometry Architecture," which are hereby incorporated by reference.

In "Quantum theory, the Church-Turing principle and the universal quantum computer," David Deutsch describes Quantum theory as a "theory of parallel interfering universes." Several embodiments of CI manipulate fundamental wave components to create constructive and destructive interference zones from which desired communication signals are created. For example, basic time-domain signals (such as pulses) may be

combined to generate complex time-domain signals. Furthermore, the frequency components of the time-domain signals may be selected to control frequency diversity and interference-rejection.

FIG. 1A and FIG. 1B illustrate one form of CI signals in which phase fronts of a plurality of CI carriers 105.A, 105.B, 105.C, and 105.D are aligned at a specific time. At other times, the carriers combine destructively resulting in undetectable signal levels. Superposition signals 110A, 110B, 110C, and 110D result from the summation of the carriers 105.A, 105.B, 105.C, and 105.D. It will be appreciated that a superposition signal may include zero or more zones of constructive interference and/or zero or more zones of destructive interference. A superposition signal may include zero or more zones of partial constructive/destructive interference. The superposition signal 110A shows a pulse envelope occurring in a predetermined time interval t_1. In the case where there is no amplitude tapering (i.e., a rectangular window) and the CI carriers are uniformly spaced in frequency, a superposition CI signal is expressed by:

$$e(t) = \sum_{n=1}^{N} e^{i[(\omega_c + n\omega_s)t + n\Delta\phi]}$$

which has a magnitude of:

$$|e(t)| = \left| \frac{\sin(N(\omega_s t + \Delta\phi)/2)}{\sin((\omega_s t + \Delta\phi)/2)} \right|.$$

The CI signals are periodic with period $1/f_s$ for odd number of carriers N and with period $2/f_s$ for even number of carriers N. The main lobe has a duration $2/Nf_s$ and each of the N-2 side lobes has a duration $1/Nf_s$. The amplitude of the l^{th} side lobe with respect to the main lobe amplitude is:

$$A(l) = \frac{1}{N \sin(\pi(l + 1/2)/N)}$$

Applying a phase shift of $n\Delta\phi_k$ to each n^{th} carrier shifts the CI envelope in time by $\Delta t = \Delta\phi_k/2\pi f_s$. Therefore, N signals can be positioned orthogonally in time. The phase shifts provide the necessary phase relationships to create the desired timing of the information signal received by at least one receiver (not shown).

The cross correlation between users is:

$$R_{cc}(\tau) = \frac{1}{2f_s}\frac{\sin(N2\pi f_s\tau/2)}{\sin(2\pi f_s\tau/2)}\cos((N-1)2\pi f_s\tau/2)$$

where τ is the time shift between envelopes. Zeros occur at k/Nf_s, $k = 1,2,...,N-1$ and at $(2k-1)/2(N-1)f_s$, $k = 1,2,...,N-1$. CI can support N orthogonal users (or channels). If additional users or signals need to be accommodated, CI provides N-1 additional positions to place signals.

Because the period and width of the pulse envelope depends on the amplitude, phase, and frequency separation of the CI carriers, the frequency of each carrier may be changed without affecting the pulse envelope as long as the amplitude, phase, and frequency separation are preserved. Thus, frequency hopping and frequency shifting of the carriers does not affect the temporal characteristics of the composite signal, such as composite signal 110.A. Providing a tapered amplitude distribution to the CI carriers broadens the main lobe width and reduces the amplitude of the side lobes.

A CI signal has a number of carrier signals that may each have a bandwidth that is less than the coherence bandwidth of the communication channel. The coherence bandwidth is the bandwidth limit in which correlated fading occurs. The total bandwidth of the CI signal preferably exceeds the coherence bandwidth.

CI signals may be spaced in frequency by large amounts to achieve a large system bandwidth relative to the coherence bandwidth. In this case, CI signals make use of the frequency diversity parameter to achieve uncorrelated fading. However, any diversity parameter or combination of diversity parameters may be used to achieve uncorrelated fading over the system bandwidth (or even between individual carriers). The system bandwidth of a group of CI carriers may be selected relative to the coherence bandwidth of one or more subchannels, such as spatial subchannels. Carriers that are closely spaced in frequency may have uncorrelated fading if they are transmitted from different locations or have different values of directivity. CI carriers transmitted from different locations may have different fades over each spatial subchannel and therefore can benefit from diversity combining at a receiver (not shown).

FIG. 1A and FIG. 1B illustrate an orthogonal time-domain spacing of $\Delta t = 2/Nf_s$ for CI pulses resulting from constructive superpositions of the carriers. Phase shifts applied to an n^{th} carrier to separate a k^{th} channel from adjacent channels is given by:

$$\phi_{kn} = \pi k n f_s(\Delta t) + \phi^o_{kn} = \pi k n/N + \phi^o_{kn}$$

where ϕ^o_{kn} is an initial phase-offset corresponding to the n^{th} carrier and the k^{th} channel.

Although FIG. 1A and FIG. 1B illustrate an in-phase superposition of carrier signals, this example can be extended to other superpositions of CI. For example, the time offset Δt (and the corresponding carrier phase shifts ϕ_{kn}) for adjacent channels may be applied to CI implementations that do not have in-phase superpositions. The time offsets Δt (and thus the phase shifts ϕ_{kn}) derived in this case are also relevant to CI implementations that process the received carriers separately. When each carrier is processed separately, phase-offset coding (in addition to the phase offsets ϕ_{kn} used to separate channels) may be used to minimize the peak of the superposition signal.

The groups of carriers 105A, 105B, and 105C have phase offsets corresponding to a pulse-width duration. However, any type of orthogonal (e.g., non-overlapping) or quasi-orthogonal (e.g., overlapping) spacing may be provided. Carrier groups having quasi-orthogonal (or non-orthogonal) spacing may be processed with multi-user (multi-channel) detection techniques or any other type of interference-suppression method.

FIG. 1A and FIG. 1B illustrate several levels of signal decomposition that reduce a complex time-domain signal into simple components. The time-domain pulses may be scaled and positioned to produce a predetermined time-domain signal indicative of an information signal, coding, and at least one transmission protocol. Multiple frequency components may be weighted to produce an information signal having predetermined time-domain characteristics. Similarly, multiple frequency components that comprise the pulses may be selected and weighted to impart predetermined characteristics to the pulses. The scale of the components selected for signal processing can be selected to provide a desired granularity for the information architecture.

Modulation of the pulses, the carriers, or both may be performed over the duration of the signals shown in FIG. 1A and FIG. 1B. Carrier modulation may be performed over a pulse-repetition period, a pulse duration, or any multiple or fraction of either. FIG. 2A and FIG. 2B illustrate how a constant weight applied to each of a plurality of carrier signals (e.g., carrier frequencies or time-offset pulses) generates a periodic code (such as a CDMA code) sequence. The code sequence may be modulated with information

symbols over a code (pulse) period or the information signal may be applied to code weights applied to the carriers during a code interval.

Information modulation applied to each group of carriers may include a relative time offset corresponding to a constant phase offset ϕ^o_{kn}. Modulation may include any modulation scheme including, but not limited to, pulse-amplitude modulation, amplitude modulation including continuous and shift key, phase modulation including continuous and shift key, frequency modulation including continuous and shift key, time-offset modulation, and any differential modulation techniques. Modulated signals may include guard bands, cyclic prefixes, codes, and the like, which are well known in the art.

FIG. 3 shows a plurality N of carrier signals that are redundantly modulated with a stream of data symbols $s_k(t)$ associated with a k^{th} channel. Each of the carriers may have a phase: $\phi_{kn} = \phi_{kn}(t) + \phi^o_{kn}$ where $\phi_{kn}(t)$ denotes a time-dependent phase and ϕ^o_{kn} denotes a constant-value initial phase. The time-dependent phase term $\phi_{kn}(t)$ can represent the relative phase variation between a plurality of carrier signals resulting from the carrier signals having different frequencies. The time-dependent phase term $\phi_{kn}(t)$ may include an impressed variation or coding to effect one or more benefits, such as reduced peak power of the superposition signal, enhanced security, improved diversity, and additional multiplexing/demultiplexing capabilities. The constant-valued terms ϕ^o_{kn} may include zero and/or non-zero values. The terms ϕ^o_{kn} may be selected to provide operational benefits, such as reduced peak power of the superposition signal, coding, and enhanced channelization.

The terms ϕ_{kn} may be defined as parts of a code $\Phi_k = \Phi(\phi,t,k,n)$ that is a function of phase, time, channel, and carrier. Thus, a redundantly modulated data stream $s_k(t)$ can be represented as having been acted upon by the code $\Phi(\phi,t,k,n)$. A received signal may include a plurality of data streams, each acted upon by various codes $\Phi(\phi,t,k,n)$. A received signal can be defined as:

$$R(t) = \sum_{k=k'}^{k''} \Phi_k s_k(t).$$

One possible method of processing an encoded data stream includes providing inverse coding to the received signal in order to decode at least one desired data stream:

$$\hat{s}_k(t) = \Phi_n^{-1} R(t).$$

Additional processing may be performed, including, but not limited to distortion compensation, multi-user detection, interference cancellation, filtering, sampling, and additional decoding.

FIG. 4A illustrates the use of different phases and time intervals applied to one set of carrier frequencies to distinguish a k^{th} channel. The set of carriers includes a plurality N of carrier frequencies, which are redundantly modulated with a stream of data symbols $s_k(t)$ associated with the k^{th} channel. However, each carrier is modulated in at least one of a plurality of time intervals, and at least some of the symbols $s_k(t)$ of the k^{th} data stream are spread across multiple time intervals. This provides both frequency diversity and time diversity to the k^{th} channel.

FIG. 4B represents a plurality of data symbols $s_1(t)$, $s_2(t)$, and $s_3(t)$ modulated onto the same carrier frequencies within the same time interval. Unlike conventional OFDM, which transmits one data bit per frequency channel per time slot, CI provides orthogonality to multiple data symbols sharing the same frequency channel and time slot. Each of the carriers having an n^{th} frequency has a phase: $\phi_{kn} = \phi_{kn}(t) + \phi^o_{kn}$, where $\phi_{kn}(t)$ denotes a time-dependent phase and ϕ^o_{kn} denotes a constant-value initial phase. In this case there are three channels, so k has values of 1, 2, and 3.

An n^{th} carrier frequency of a k^{th} channel may have a time offset t_{kn} where:

$$t_{kn} = t_{kn}(t) + t^o_{kn}.$$

The term $t_{kn}(t)$ denotes a time-varying time offset (or interval start time) and t^o_{kn} denotes a constant-value initial time offset. The time-varying term $t_{kn}(t)$ may represent a plurality of time intervals that are uniformly spaced (e.g., $t_{kn}(t) = m\Delta t_{kn}$, where $m = 0,1,2,\ldots$, and Δt_{kn} is some constant time interval) or non-uniformly spaced. A k^{th} channel may have a phase code Φ_k and/or a time code T_k wherein phase values ϕ_{kn} and time values t_{kn} are predetermined or adapted with respect to at least one performance objective, such as reducing peak power, improving a signal-to-interference plus noise ratio, enhancing diversity, providing security, reducing a bit-error rate, etc.

FIG. 5 illustrates a plurality of time-offset carrier signals having different initial time offsets t^o_{kn}. Data streams (or any other type of information signal) may be redundantly modulated onto the time-offset carrier signals, either directly or with respect

to some spreading code corresponding to time-offsets, phase-offsets, carrier frequency, phase, or the like. The time intervals may have uniform or non-uniform duration. Each time interval typically has a duration of about a symbol interval. However, the time intervals may have durations that are longer or shorter. The time intervals may overlap, adjoin each other, or they may be separated.

One method for demultiplexing a k^{th} channel from a received signal includes compensating for coded time offsets and/or phase offsets and then processing the signal components separately. Each carrier may be processed separately by a matched filter and decision system. Another method involves processing superpositions of the compensated components. The superpositions may be processed by any well-known detection technique, such as matched filtering, down-conversion, envelope detection, demodulation, decoding, etc. Another method for demultiplexing a k^{th} channel from a received signal includes matched filtering of the components of the coded signal and/or the superposition of the components.

FIG. 6A illustrates a plurality of carriers corresponding to a plurality M of subchannels each having a plurality N of carriers. The carriers shown are spaced in frequency by a constant amount f_s. The total bandwidth of the channel is proportional to NMf_S. The bandwidth of a subchannel is related to Nf_S.

FIG. 6B illustrates a superposition of three subchannels that occupy the same time-domain space. However, the separability of the time-domain signals into their frequency components provides orthogonality to the signals (and the subchannels). A consideration of FIG. 6A and FIG. 6B illustrates how a wideband signal can be separated into smaller-bandwidth subchannels that are easier to process in the time domain. The duration of each data symbol in a subchannel is M times the duration of data symbols in the total channel. However, each subchannel occupies almost the full bandwidth occupied by the total channel, and thus, derives the same frequency-diversity benefits as the total channel.

FIG. 7 illustrates how time offsets may be used simultaneously with carrier-defined subchannels. Modulated carriers corresponding to each subchannel may be decoded relative to a time-spreading code and processed separately or combined to produce a superposition signal. The subchannels may employ other types of encoding

(such as phase encoding) as well. Phase decoding, as well as time-offset decoding and carrier-frequency selection may be performed to separate subchannels and facilitate processing of the data streams modulated on the subchannels.

Subchannels $s_1(t)$, $s_2(t)$, and $s_M(t)$ are shown having carrier components occurring in three different time intervals. An m^{th} communication channel may be determined by an m^{th} code sequence of carrier frequencies and time slots. For example, a symbol value for subchannel $s_1(t)$ is modulated onto the following carrier frequency-time interval pairs: f_1:t_2, f_{M+1}:t_0, and $f_{(N-1)M+1}$:t_1. More than one subchannel may use the same set of carrier frequencies and time intervals if different phases are provided. Although FIG. 7 shows uniformly spaced frequencies and time slots, non-uniform spacing may be employed.

A subchannel data stream encoded according to one or more of the techniques described in the specification may be processed at a receiver using one or more decoding techniques. A receiver may shift a plurality of its carriers with respect to a decoding sequence corresponding to one or more transmitted channels. The decoded signals may be processed separately or combined to create a superposition signal. Alternatively, the receiver may generate a reference signal based on the code(s) of one or more channels and use the reference signal in a matched filter or equivalent receiver system to process the received signals.

3. CI Transmission Methods

FIG. 8A illustrates a CI transmission method. A carrier-generation step 801 provides a plurality of carrier frequencies. A phase-application step 802 provides a predetermined phase relationship to the carriers. The phase-application step 802 may be combined with the carrier-generation step 801. A modulation step 803 redundantly modulates the carrier frequencies with at least one data stream to generate at least one set of modulated carriers. An optional multiplexing step 804 multiplexes a plurality of sets of modulated carriers. An optional coupling step 805 couples the modulated carriers into at least one communication channel.

The carriers may be generated using Fourier transform techniques (e.g., FFT, IFFT, DFT, IDFT), pulse generation, non-linear combining of reference signals, harmonic generation, frequency-shifted feedback, or any technique that generates a multicarrier signal having a plurality of frequencies observed by a receiver. The

individual carriers may be provided with predetermined amplitudes that are constant or time varying. Each carrier may include one or more subcarriers. The carriers may be provided with one or more phase relationships during the carrier generation step 801 or in at least one separate phase-application step 802. Carriers may be phase shifted, delayed, modulated, or otherwise acted upon to provide at least one set of phase relationships.

The carriers may be modulated with information prior to, during, or after the phase-application step 802. Alternatively, information modulation may be applied during the carrier-generation step 801. For example, a periodic-pulse generator may be modulated such that generated pulses (and thus, the uniformly spaced carrier signals) are modulated with at least one data stream.

FIG. 8B shows steps of a CI transmission method. A carrier-generation step 801 provides a plurality of carrier frequencies. A modulation step 803 redundantly modulates the carrier frequencies with at least one data stream to generate at least one set of modulated carriers. An optional multiplexing step 804 multiplexes a plurality of sets of modulated carriers. An optional coupling step 805 couples the modulated carriers into at least one communication channel.

The modulation step 803 modulates the carriers with respect to at least one predetermined set of diversity parameters, such as carrier frequencies, times, and phases. Modulation may be performed with respect to one or more codes including, but not limited to spreading codes, security codes, multiple-access codes, bit-loading codes, and power-reduction codes.

FIG. 8C illustrates a method of generating CI signals. A carrier-generation step 801 provides a plurality of carrier frequencies. A channelizing step 806 establishes at least one multiple-access subchannel by performing at least one of the following functions: 1) selecting a plurality of the carrier frequencies, 2) selecting or impressing a carrier phase relationship, and 3) selecting or impressing a set of carrier time intervals. A modulation step 803 redundantly modulates the carriers in each subchannel with at least one data stream.

4. CI Reception Methods

FIG. 9A illustrates a method of receiving and processing CI signals. A receiving step 901 includes receiving at least one CI signal from a communication channel. A

decoding step 902 describes any of a set of methods that may be used to decode a received signal encoded explicitly by one or more transmission sources and/or encoded incidentally (e.g., propagation effects, distortion) by the communication channel. A processing step 904 processes the decoded signals to extract one or more data streams.

The receiving step 901 may include any method of coupling (or otherwise being responsive to) at least one transmitted signal into a receiver apparatus. The decoding step 902 includes compensating for at least one diversity parameter value (such as phase, time, polarization, amplitude) adjusted with respect to at least one explicit code and/or incidental code. Results of decoding the received signal(s) may include a superposition signal having at least one desired time-domain characteristic.

The processing step 904 translates the decoded signal(s) into one or more data streams. One or more superposition signals may be processed via any well-known reception techniques including matched filtering, down-conversion and baseband detection, envelope detection, constellation processing, direct detection, decision feedback, etc. A plurality of the received carrier signals may be processed separately, such as by using matched filtering. Both superposition processing and multicarrier processing may be performed on the same set of received signals. The processing step may include front-end processing (i.e., during the receiving step 901) and/or intermediate processing (i.e., during the decoding step 902). The processing step 904 may include one or more additional processing capabilities including, but not limited to filtering, sampling, A-to-D conversion, D-to-A conversion, amplification, equalization, decision, multi-user detection, error correction, channel estimation, adaptation to training symbols, weighting, delaying, interference cancellation, integration, comparison, decoding, array processing, diversity combining, and optimization techniques.

FIG. 9B illustrates a method of receiving CI signals. A receiving step 901 includes receiving at least one CI signal from a communication channel. A reference-signal generation step 903 includes generating a coded reference signal with respect to explicit and/or incidental coding applied to at least one received signal. A processing step 904 processes the reference and received signals to extract one or more data streams. A superposition of the reference and received signals may be processed via matched filtering. Similarly, carrier components of the reference and received signals may be

processed via matched filtering. The processing step 904 may include one or more additional processing steps as described with respect to FIG. 9A.

The transmission and reception techniques presented herein may be applied to any CI architecture. The steps comprising the transmission and reception methods may be adapted to the type of CI architecture employed. For example, spatial carrier interferometry may be performed using spatially separated antennas to enhance diversity benefits of CI signals. Appropriate array-processing techniques may be integrated into CI transmission and/or reception methods. Additional CI architectures (such as described herein and in some of the cited references) may be supported using appropriate adaptations to the methods and systems described herein.

The CI transmission and/or reception methods described herein may include additional processing to shape transmitted and/or receive signals to provide desired time-domain characteristics. For example, the peak-to-mean-power of a multicarrier superposition signal may be reduced by appropriate selections of carrier phases and frequencies. Time-domain characteristics of conventional transmission protocols may be provided by appropriate selections of carrier amplitudes and phases. Consequently, time-domain shaping considerations may be incorporated into the CI methods and systems described herein. Despite the adaptation of CI to time-domain processing, CI signals may be generated and received using the basic methods and systems described herein.

5. CI Transmitters

FIG. 10 shows a functional diagram for a system adapted to generate CI signals. An information signal $s_k(t)$ (from an input data source 101) intended for a k^{th} user is modulated onto N carriers in a first modulation process 204A. In a second modulation process 204B, complex weights are applied to the modulated carriers. The complex weights may include phase shifts (or delays). Unlike a chip sequence in MC-CDMA (which uses binary values, such as ±1), CI signals use incremental (n) phase-shift values, such as $e^{in\Delta\phi}{}_k$. The process of providing a uniform delay to each of the carriers applies incremental phase-shift values (such as $e^{in\Delta\phi}{}_k$) to the carriers. In an optional third modulation process 204C, additional weights a_{nk} are applied to the carriers. The weights a_{nk} may include windowing weights, channel-compensation values, code values, and/or weight values that facilitate signal separation by cancellation or constellation methods at

a receiver (not shown). The weights a_{nk} may include a diversity operation $P(q)$. The modulation processes 204A, 204B, and 204C may be performed in any order and may be combined. The carriers are combined 210, optionally up-converted 215, and coupled 250 into a communication channel 99.

A CI transmission system shown in FIG. 11 includes a multicarrier generator 220, an information source 214, a modulator 222, a superposition controller 221 (such as a phase controller), and an output coupler 223. The multicarrier generator 220 produces a multicarrier signal that is modulated with at least one information signal from the information source 214. Alternatively, the information source 214 may provide the information signal to the multicarrier generator 220, which impresses the information signal onto the multicarrier signal. The superposition controller 221 provides a predetermined relationship (such as a particular relative phase relationship) to the multicarrier signal that causes a superposition of the carriers to have a predetermined relationship (such as a particular time offset). The superposition controller 221 may be integrated into the multicarrier generator 220 such that the multicarrier signals are generated with the predetermined relationship. In another embodiment, the superposition controller 221 may be integrated into the modulator 222. In a further embodiment, the function of the superposition controller 221 may be enhanced by dispersion characteristics of a communication channel (not shown).

The multicarrier generator 220 may include any type of system that generates a plurality of carrier signals characterized by different values of at least one diversity parameter. For example, a multi-frequency carrier generator may include one or more of the following: a pulse generator (such as a switching system), a plurality of local oscillators (LOs), a mode-locked system, a frequency-shifted feedback system, an FFT circuit, an IFFT circuit, a nonlinear system capable of generating a harmonic and/or inter-modulation products, and any type of wideband generator coupled to a filter bank. Additional embodiments of CI transmission systems are shown in FIG. 12A and FIG. 12B.

FIG. 12C shows a procedural description of a CI-CDMA transmitter. CI-pulse generation 410 involves generating 411 a multicarrier signal that is combined 412 into a modulated superposition pulse having a pulse shape $h(t)$:

$$|h(t)| = \left| A \cdot \frac{\sin(\frac{1}{2} N 2\pi\Delta f t)}{\sin(\frac{1}{2} 2\pi\Delta f t)} \right|$$

To transmit an i^{th} chip, an i^{th} value of a spreading sequence $\beta_i^{(k)}$ (where $\beta_i^{(k)} = \pm 1$) is modulated by a CI pulse-shape filter 413, $h(t-iT_c)$. User k's transmission signal is:

$$s_k(t) = b_k \sum_{i=0}^{N-1} \beta_i^{(k)} \sum_{n=0}^{N-1} \cos(2\pi\omega_c t + 2\pi n \Delta f (t - iTc)) g(t)$$

where b_k is a data bit for the k^{th} user and $g(t)$ is a unit-amplitude rectangular waveform having a duration of T_s. The expression for $s_k(t)$ can be written as:

$$s_k(t) = b_k \sum_{i=0}^{N-1} \beta_i^{(k)} \sum_{n=0}^{N-1} \cos(2\pi\omega_c t + 2\pi n \Delta f t - ni2\pi / N) g(t)$$

where $T_c\Delta f = T_c/T_s = 1/N$. The total transmitted signal for K users is:

$$S(t) = \sum_{k=0}^{K-1} b_k \sum_{i=0}^{N-1} \beta_i^{(k)} \sum_{n=0}^{N-1} \cos(2\pi\omega_c t + 2\pi n \Delta f t - ni2\pi / N) g(t)$$

The time-shifted pulses are orthogonal to each other:

$$\int_0^{Ts} h(t - pTc) h(t - qTc) dt = 0 \qquad (p \neq q).$$

The chips $h(t-iT_c)$ can be created by phase offsetting each n^{th} carrier of the pulse $h(t)$ by $2\pi n/N$.

6. CI Receivers

A basic CI receiver for a k^{th} user is shown in FIG. 13A. Signal components received from a communication channel (not shown) have various values of amplitude A_{kn} and phase ϕ_{kn}. These values may differ between received signal components due to amplitude and phase profiles of the transmitted signals, effects of the communication channel on the amplitudes A_{kn} and the relative phases ϕ_{kn} of the transmitted signals, and any effects caused by reception. Transmission and/or reception may be controlled to provide predetermined relative amplitudes A_{kn} and/or phases ϕ_{kn} to the received signals. The relative values of amplitude A_{kn} and phase ϕ_{kn} of each received component are taken into consideration when providing a cross-correlation signal to each component. For example, relative values A'_{kn} and ϕ'_{kn} of the cross-correlation signal components may be matched to the relative values A_{kn} and ϕ_{kn} of the received signal components.

Signal components received from the communication channel (not shown) may have sparsely distributed values of frequency f_{kn} and/or and phase ϕ_{kn}. The term sparsely distributed, as used herein, refers to carrier frequencies f_{kn} and/or phases ϕ_{kn} that have a plurality N of values where $N > 2$. The carrier frequencies f_{kn} and/or phases ϕ_{kn} may be uniformly spaced or non-uniformly spaced. Sparsely distributed values may be characterized by a random or chaotic distribution. Various types of mathematical distributions may be used to characterize sparsely distributed values.

A CI receiver for a k^{th} user is shown in FIG. 13B. CI signals are coupled out of a communication channel (not shown) by a coupler 301. Information signals are extracted from each carrier by a plurality N of correlators 305a to 305N. The correlators 305a to 305N may include a filter bank (not shown). In this case, the correlators 305a to 305N project the received signals onto the orthonormal basis of the transmitted signals. The constant-phase value $\Delta\phi_k$ for a k^{th} user equals $2\pi k/N$. Correlators (such as the correlators 305a to 305N) may compensate for channel distortion and/or addressing. The correlators 305a to 305N may apply windowing and/or other filtering processes to the received signals.

Correlation signals output from the correlators 305a to 305N may be integrated by a plurality of integrators 314a to 314N over a symbol interval T_s before being combined in a combiner 319. A decision device 355 detects the combined signals. The decision device 355 may be part of the combiner 319. The decision device 355 may perform multi-user detection or multi-channel detection and may perform any combination of cancellation and constellation processes to determine the value(s) of received signal(s).

FIG. 14A shows an embodiment of CI receiver configured to operate with DS-CDMA. A plurality N of chip receivers 405a to 405N process each time-domain chip separately before combining the processed chips in a DS-CDMA combiner 415. An output signal from the combiner 415 is processed by a decision system 417.

FIG. 14B shows an embodiment of one of the chip receivers 405n. Each carrier of a chip is down-converted to baseband by a down-converter system 407a to 407(N-1). Each baseband signal is processed (e.g., integrated or sampled) over a symbol duration T_s by an integrator (or sampling circuit) 414a to 414(N-1). The symbol duration T_s is typically

substantially longer than the chip duration T_c. The processed baseband signals are combined in a combiner 416.

In a Rayleigh frequency-selective slow fading channel, each multi-frequency carrier comprising the CI-CDMA signal experiences a unique flat fade. The received signal is represented as:

$$r(t) = \sum_{k=0}^{K-1} b_k \sum_{i=0}^{N-1} \beta_i^{(k)} \sum_{n=0}^{N-1} \alpha_n \cos(2\pi\omega_c t + 2\pi n\Delta f t - ni2\pi / N + \phi_n) g(t) + n(t)$$

where α_n is a gain and φ_n is a phase offset of an n^{th} carrier of a CI pulse resulting from fading, and $n(t)$ is additive white Gaussian noise. Assuming exact phase synchronization, a decision variable resulting from a received multi-frequency carrier component of each received chip is expressed by:

$$r_{m,n} = \sum_{k=0}^{K-1} b_k \sum_{i=0}^{N-1} \beta_i^{(k)} \alpha_n \cos(nm2\pi / N - ni2\pi / N) + n_{m,n}$$

The CI-CDMA receiver for an l^{th} user detects an m^{th} chip at each carrier, as shown in FIG 14B. The m^{th} chip is separated into its N carrier components. Each chip and each carrier contribute a decision variable $r_{m,n}$ corresponding to

$$r_{m,n} = b_l \beta_m^{(l)} \alpha_n + b_l \alpha_n \sum_{\substack{i=0 \\ i \neq m}}^{N-1} \beta_i^{(l)} \cos(nm2\pi / N - ni2\pi / N)$$
$$+ \sum_{\substack{k=0 \\ k \neq l}}^{K-1} b_k \beta_m^{(k)} \alpha_n + \sum_{\substack{k=0 \\ k \neq l}}^{K-1} b_k \alpha_n \sum_{\substack{i=0 \\ i \neq m}}^{N-1} \beta_i^{(k)} \cos(nm2\pi / N - ni2\pi / N) + n_{m,n}$$

The first term represents the contribution from user l, carrier n, and chip m. The second term represents the other N-1 chips of the user l. The third term represents interference due to the m^{th} chip of other users. The fourth term represents interference from user l's other N-1 chips. The fifth term is a zero-mean Gaussian random variable with variance $\sigma_n^2=N_0/2$. The terms, $n_{m,n}$, are correlated across chips, but not across carriers.

The covariance matrix of the vector noise $(n_{0,n}, n_{1,n}, n_{2,n}, \ldots n_{N-1,n})$ corresponding to a fixed carrier number n and a variable chip number m is:

$$C_n = \frac{N_0}{2}\begin{pmatrix} 1 & \cos(2\pi/N) & \cos(2*2\pi/N)\bullet\bullet\bullet\cos((N-1)*2\pi/N) \\ \cos(2\pi/N) & 1 & \cos(2\pi/N) \quad \bullet\bullet\bullet\cos((N-2)*2\pi/N) \\ & & \bullet \\ & & \bullet \\ & & \bullet \\ \cos((N-1)*2\pi/N) & & \bullet\bullet\bullet \qquad 1 \end{pmatrix}$$

Multi-frequency carrier combining is used to combine the $r_{m,n}$'s across the carriers, as indicated by the functionality shown in FIG. 14B. This results in frequency-diversity benefits when recreating each chip and removal of the second and fourth interference terms (which represent inter-chip interference).

Orthogonality Restoring Combining (ORC) may be used to remove the second and fourth interference terms. Each $r_{m,n}$ term is scaled by α_n and summed over n to provide a decision variable R_m for an m^{th} chip:

$$R_m = \sum_{n=0}^{N-1} r_{m,n} \bullet 1/\alpha_n$$

However, ORC can result in substantial noise enhancement. Thus, ORC is most suitable for low-noise conditions (i.e., high signal-to-noise).

Equal Gain Combining (EGC) is preferable for low signal-to-noise. EGC combines the N carrier terms for the m^{th} chip according to:

$$R_m = \sum_{n=0}^{N-1} r_{m,n}$$

Minimum Mean Square Error Combining (MMSEC) minimizes the second term and the fourth term and optimizes frequency diversity while minimizing the noise. Using multi-carrier MMSEC provides a decision variable R_m for an m^{th} chip:

$$R_m = \sum_{n=0}^{N-1} r_{m,n} \bullet (\alpha_n / (K\alpha_n{}^2 + N_0))$$

A final decision variable D_l for user l results from a typical DS-CDMA combining technique across chips, which eliminates multi-user interference. Each chip's decision variable R_m is multiplied by an m^{th} spreading code $\beta_m^{(l)}$ and combined:

$$D_l = \sum_{m=0}^{N-1} R_m \beta_m^{(l)}$$

The orthogonal cross-correlation between spreading codes of different users minimizes the multi-user interference.

7. Duplexing CI Signals

CI enables any type of duplexing to separate transmitted and received signals. The multicarrier structure of CI enables frequency-division duplexing (FDD). Even spread-spectrum protocols, such as CDMA, may benefit from CI-enabled FDD without the reduced frequency-diversity benefits inherent in conventional FDD. Time-division duplexing (TDD) may be provided with respect to time-domain characteristics of multicarrier superposition signals and/or time-offset coding applied to CI.

CI phase-division duplexing (CI-PDD) describes the separation of multicarrier signals having different phase spaces. Transmitted and received CI signals may be separated due to the orthogonality (or quasi-orthogonality) of the phase spaces. One application of CI-PDD involves applying a matched filter to desired received components to remove the effects of transmitted signals (as well as undesired received signals).

Cancellation may be used to electromagnetically isolate receivers from transmitters that use the same time and frequency bands for transmission. Some of these cancellation techniques are described in PCT Appl. No. PCT/US94/08247.

Polarization division duplexing includes orthogonal polarization duplexing (OPDD) and quasi-orthogonal polarization duplexing (QPDD). OPDD is a well-known method of using orthogonal polarizations to double channel capacity. For example, linear polarized antennas may be oriented perpendicular to each other to provide isolation. OPDD involves using known ratios of co-channel interference to cancel the interfering signals.

Coherence time division duplexing (CTDD) may be used to remove interfering signals having different time offsets. In CTDD An information signal modulated onto a wideband signal is transmitted along with a decode signal. A predetermined time offset between the modulated and decode signals allows a receiver to extract the information signal without any knowledge of the wideband signal.

Orthogonal CTDD pertains to time-offset differences between transmitted and received signals that exceed the inverse of the coherence bandwidth of the signals. Quasi-orthogonal CTDD occurs when the time-offset differences between desired receive

signals and interfering transmissions do not exceed the inverse coherence bandwidth of the signals. This causes overlap interference in the matched filter output. The overlap interference may be removed via cancellation.

8. Beam Forming with other Diversity Parameters

In conventional antenna arrays, spatially separated antennas receive different signal phases (or delays) of non-normally incident radiation. The spatial separation between the antennas provides the array with unique phase responses that enable beam forming via adaptive or deterministic weight-and-sum techniques. Environmental effects (such as path loss, multipath fading, scattering, diffraction, etc.) provide complex-amplitude variations (spatial gain distributions) that enable novel array-processing methods described in Applicant's co-pending applications and patents already included by reference. Similarly, individual antennas or groups of antennas may be provided with unique directional gain patterns such that the individual antennas or groups of antennas have different signal-amplitude responses (as well as different phase responses) to incident radiation.

FIG. 15 shows a spatial-interferometry receiver that includes multiple stages of interference cancellers. The receiver includes an array of receiver elements 141.1 to 141.M. A first processing stage 161 includes a plurality of beam formers 161.1 to 161.M', each having directional-gain responsiveness indicated by one of a plurality of directional beam patterns (not shown). A second processing stage 103 may include a weight-and-sum canceller that removes or separates interfering signals in the beam patterns.

The spatial-interferometry receiver may be part of a communication network that overlays multiple channels in a predetermined frequency band, causing interference at the receiver. Interfering signals that are external to the network (i.e., uncontrollable interference) may be nulled in the first processing stage 161. Intersymbol interference may be nulled in the first processing stage 161. Desired interfering signals coupled into the second processing stage 103 may be separated using one or more interference-cancellation and/or interference-suppression techniques (such as weight-and-sum, matrix diagonalization, optimization, etc.). Interference cancellation and/or suppression may be

performed to optimize some measurable parameter, such as signal-to-noise ratio, signal-to-noise-plus interference ratio, signal-to-interference ratio, bit-error rate, etc.

A plurality of receivers adapted to have different responsiveness to different values of at least one diversity parameter can be processed with respect to the diversity-parameter space(s) in which the receivers are adapted. For example, it is well known that various forms of spatial processing may be performed on signals received from spatially separated antennas. However, these spatial-processing techniques may be applied to other diversity-parameter spaces and subspaces.

The simplest beam-forming technique involves a fixed beam pattern approach (e.g., sectorization). Adaptive techniques may be applied to beam forming to compensate for changing conditions between transmitters and receivers. Angle of arrival and angle of departure may be determined based on any of several optimization techniques, such as maximizing radiation projected to at least one desired user and/or minimizing radiation projected at undesired users. Several well-known algorithms are suitable for estimating an angle-of-arrival spectrum, such as MUSIC, ESPRIT, and WSF. Feedback techniques may be employed to enhance adaptive beam forming. Open-loop and closed-loop feedback methods are commonly used to optimize beam forming. Another variant of beam forming includes analyzing received signals to assist in generating a transmit beam pattern.

Many beam-forming techniques employ channel estimation to account for multipath effects. In theory, an optimum transmit beam pattern may be developed based on certain statistical properties of received signals. Blind adaptive methods may be used to optimize one or more performance measurements. For example, a beam-pattern weight matrix may be adapted relative to 1) at least one covariance matrix characterizing a communication channel between a plurality of transmitters and at least one desired receiver, and 2) at least one covariance matrix characterizing signal propagation to at least one undesired user.

Optimization processing techniques typically are designed to maximize the signal to interference-plus-noise ratio. This may be accomplished by adjusting the array weights to minimize the mean squared error of an output signal relative to a known reference

signal. The reference signal may be a coherent or constant-modulus reference. Weights may be adapted relative to a noise-plus-interference correlation matrix.

Well-known adaptive beam-forming techniques that can also be applied to non-spatial diversity-parameter spaces are described in the following publications, which are incorporated by reference: U.S. Pat. No. 5,875,216, U.S. Pat. No. 6,101,399, U.S. Pat. No. 4,734,701, U.S. Pat. No. 5,414,699, U.S. Pat. No. 4,931,977, U.S. Pat. No. 5,414,699, U.S. Pat. No. 5,828,658, U.S. Pat. No. 5,796,779, U.S. Pat. No. 5,471,647, U.S. Pat. No. 5,592,490, and U.S. Pat No. 6,144,711.

FIG. 16 shows a relationship between a pair of orthogonal linear-polarization directions. These directions may correspond to linear polarizations of transmitted signals, linear transmitter polarization, and/or linear receiver polarization. These directions may correspond to components of a circular-polarized signal, a circular-polarized transmitter, and/or a circular polarized receiver. The polarizations can be related to arbitrary *x*- and *y*-axes:

$$P_1 = \cos\vartheta\hat{x} + \sin\vartheta\hat{y}$$

$$P_2 = -\sin\vartheta\hat{x} + \cos\vartheta\hat{y}$$

If θ is time variant (e.g., $\theta = \omega t$), then the vector sum of P_1 and P_2 represents a circular-polarized signal having an angular velocity of ω. The intensity of a polarized signal received by a polarized receiver having an angular offset of φ is indicated by:

$$I = \cos\vartheta\cos(\vartheta+\varphi) + \sin\vartheta\sin(\vartheta+\varphi)$$

If θ is a constant (linear polarization) or time variant with a constant angular velocity ω (circular polarization), the intensity I reduces to: $I = \cos\varphi$. Thus, perpendicular polarizations, whether linear or circular (e.g., elliptical), are orthogonal.

FIG. 17 shows a "beam pattern" for a polarized receiver. The beam pattern is a polar plot of intensity versus polarization offset φ relative to the receiver's polarization. The receiver may have linear or circular polarization. If the receiver is elliptically polarized, the intensity changes with respect to the magnitudes of the minor and major axes as well as with respect to the polarization offset φ. A first signal $s_1(t)$ having an amplitude A_1 and a phase ϕ_1 is received with a gain g_1 corresponding to the signal's $s_1(t)$ relative polarization offset φ_1. Similarly, a second signal $s_2(t)$ having an amplitude A_2 and

a phase ϕ_2 is received with a gain g_2 corresponding to the signal's $s_2(t)$ relative polarization offset φ_2. The received signal is a superposition of the signals $s_1(t)$ and $s_2(t)$:

$$R = g_1 A_1(\phi_1) + g_2 A_2(\phi_2)$$

The gains g_1 and g_2 may be selected to cancel the received signals $s_1(t)$ and $s_2(t)$. For example, if signals $s_1(t)$ and $s_2(t)$ are multipath components of a single transmission or otherwise redundant transmissions, then the gains may be selected to weight the complex amplitudes $A_1(\phi_1)$ and $A_2(\phi_2)$ such that they cancel.

FIG. 18A shows two beam patterns wherein each beam pattern corresponds to one of two polarized transceivers. A first transceiver has a null in a direction φ_1 and a second transceiver has a null in a direction φ_2. The first transceiver has a gain of $g_1(\varphi)$ and the second transceiver has a gain of $g_2(\varphi)$ in a polarization direction φ. The polarization offset φ may be nulled if appropriate weights w_1 and w_2 are applied to the beam patterns:

$$R = w_1 g_1(\varphi) A(\phi) + w_2 g_2(\varphi) A(\phi) = 0$$

FIG. 18B shows a beam pattern resulting from the weighted sum of the beam patterns shown in FIG. 18A.

FIG. 19A shows a pair of linear-polarized directions having a relative polarization offset of φ. A polarized direction is meant to convey a polarization of a transmitter (or a transmitted signal) with respect to a polarization of a receiver. FIG. 19B shows a pair of circular (or elliptical) polarized directions having identical angular velocities ω and a polarization offset of φ. Beam patterns shown in FIG. 18A and FIG. 18B are applicable to linear polarized or circular (or elliptical) polarized directions, such as shown in FIG. 19A and FIG. 19B, respectively.

In FIG. 19C, a pair of circular (or elliptical) polarized directions having different angular velocities ω and ω' ($\omega \neq \omega'$) provides a relative angular velocity: $\Delta\omega = \omega - \omega'$. FIG. 20A shows a plot of a polarization offset $\varphi(t)$ over a period $T = 1/\Delta\omega$ resulting from a pair of circular polarized directions having different angular velocities ω and ω'. FIG. 20B shows a pair of polarization offsets $\varphi(t)$ and $\varphi'(t)$ over a period $T = 1/\Delta\omega$ resulting from two pairs of circular polarized directions having different initial polarization offsets φ_o and φ'_o. Each pair of polarization offsets $\varphi(t)$ and $\varphi'(t)$ have the same difference in

angular velocities $\Delta\omega$. The pair of polarization offsets $\varphi(t)$ and $\varphi'(t)$ have a non-zero relative phase (difference in initial polarization offsets φ_o and φ'_o).

Periodic polarization offsets (such as shown in FIG. 20A and FIG. 20B) may be processed in beam-forming operations. For example, interfering polarization offsets over one or more time intervals may be weighted in multi-user detection and/or interference-cancellation techniques. Periodic polarization offsets may be processed using processing techniques that exploit the periodic nature of the offsets. For example, the processing techniques may include circular convolutions, discreet Fourier transforms, and/or other methods that make use of the periodic nature of the offsets. Similarly, other periodic diversity-parameter spaces (such as phase spaces, under-sampling spaces, and/or any other diversity-parameter space in which time-dependent functions provide a periodic variation in signal detection) may be exploited.

FIG. 20C shows a plurality N of polarization offsets $\varphi(f,t)$ having different frequencies f_i. In this case, the polarization offsets $\varphi(f,t)$ are locked in phase at time t_o. The superposition of receiver responses to the polarization offsets $\varphi(f,t)$ is substantially non-zero during a narrow time-domain interval t_p. If the frequencies f_i are uniformly spaced by an amount of f_s, then the interval t_p is $1/(f_sN)$, which occurs periodically with period $T_s = 1/f_s$. Thus, periodic signal-polarization (as well as other periodic diversity parameters and periodic responses) may be processed with respect to methods described and referenced herein with respect to CI.

FIG. 21A shows a pair of gain distributions $g_1(f)$ and $g_2(f)$ in frequency space for a pair of receivers having different frequency responses. The gain distributions $g_1(f)$ and $g_2(f)$ may be regarded as beam patterns in frequency space for each of the receivers. The combination of the beam patterns (preferably after weighting) may be regarded as an array beam pattern expressed by:

$$G(f) = w_1 g_1(f) + w_2 g_2(f)$$

where w_1 and w_2 are complex weights that may have any value including unity.

A desired signal $s_1(f_1)$ centered at a frequency f_1 may be observed with minimal interference from an interfering signal $s_2(f_2)$ centered at frequency f_2 provided that appropriate beam-forming weights are applied to the receiver responses $g_1(f)$ and $g_2(f)$. Thus, beam forming in frequency space can reduce the complexity of filtering.

FIG. 21B shows a pair of gain distributions $g_1(\chi)$ and $g_2(\chi)$ in a non-spatial χ-space for a pair of receivers having different responses in the χ diversity-parameter space. Beam forming may be performed in any non-spatial diversity-parameter space including, but not limited to any combination of frequency, phase, time, correlation, polarization, code, subspace, mode, and any time-dependent diversity parameter thereof.

9. Beam Forming in Phase Space and its Relation to CI

In OFDM, as in some implementations of CI, information symbols are modulated onto uniformly spaced-in-frequency carriers. The duration T_s of a symbol interval is typically equal to $1/f_s$, where f_s is the frequency spacing between carriers. The frequency spacing f_s may be different than frequency spacings of carriers in one or more subchannels.

FIG. 22A illustrates a correlation between a pair of carrier signals having similar frequency f and phase ϕ. The correlation of the carriers is performed over a time interval of $T_s = 1/f_s$ and the correlation is normalized (=1). The correlation procedure performed in the mathematical analysis related to some of the figures shown in the specification included a sum of terms resulting from an inner product of vectors representing each carrier.

FIG. 22B illustrates how a correlation between carrier signals having a frequency spacing that is an integer multiple of f_s is equal to zero over the interval T_s. A first carrier having a frequency f_1 and phase ϕ_1 is correlated with a second carrier having a frequency f_2 and phase ϕ_2. Thus, carriers that are spaced by an integer multiple of f_s represent a set of orthogonal channels.

FIG. 22C illustrates how carriers having similar frequencies but different phases have a normalized correlation value between 1 and −1. In the case of a $\pi/2$ phase difference (shown in FIG. 22D), the carriers are defined as in phase and quadrature with respect to each other and correlate to zero. Except for $\pi/2$ relative phase encoding, differential phase encoding does not permit orthogonal channels.

FIG. 23 shows a functional diagram of a phase-space beam-forming system. Multiple data streams $s_n(t)$ are modulated onto a plurality of carriers. Each carrier is defined by a particular frequency f_1 and a unique phase ϕ_n. In this case the carriers have similar frequencies f_1. The carriers are processed in a plurality of correlators 237.1 to

237.N. Each correlator 237.1 to 237.N generates a desired signal and one or more interfering signals. The correlator outputs are coupled to a beam-forming system 238 (such as a weight-and-sum circuit, a multi-user detector, or any type of beam former) that substantially separates the desired signals from interference. The beam-forming system 238 may provide phase-space processing for a plurality of carrier frequencies. The beam-forming system 238 may incorporate additional diversity parameters into its interference-suppression and/or beam-forming processes, such as frequency, space, time, polarization, etc.

OFDM converts a serial data stream into parallel data streams wherein different symbols are transmitted over different carrier frequencies. Orthogonality of the carrier frequencies ensures that data symbols on different carriers do not interfere with each other during reception. Thus, OFDM converts a serial data stream into a plurality of orthogonal frequency channels.

CI differs from OFDM in that CI enables redundant modulation of a plurality of carriers with one or more data streams s_n(t) without sacrificing system capacity. CI differs from OFDM in that it uses interference (rather than frequency separation) between different carriers to achieve orthogonal subchannels. Unlike a substantially orthogonalizing procedure (e.g., OFDM), CI creates a destructive superposition of interfering signal components to cancel interference and separate subchannels.

FIG. 24 shows a superposition of 11 phase-locked carrier frequencies modulated with an information symbol $s_n(t_o)$. The application of the data stream $s_n(t)$ to a set of carriers is defined by a phase-space function $\Phi_n(f_i, \phi_j)$ that indicates frequencies f_i and phases ϕ_j of the carriers modulated with the data stream $s_n(t)$:

$$s_n(t)\Phi_n(f_i, \phi_j) = s_n(t, f_1, \phi_1) + s_n(t, f_2, \phi_2) + \ldots + s_n(t, f_{N'}, \phi_{N'})$$

The frequency separation of the carriers is f_s and the time interval shown is $T_s = 1/f_s$. The superposition signal is shown having a pulse width of $t_p = T_s/N$, where $N = 11$ is the number of carriers. The pulse is centered at time-index 0, where each of the carriers has a maximum. It will be appreciated that carrier phases, amplitudes, and/or frequencies may be selected to reduce the peak of the superposition signal. The carrier signals may be processed in a receiver using either or both the superposition signal and the set of individual carrier signals.

In single-carrier and non-redundantly modulated multicarrier systems, differential phase encoding (except for $\pi/2$ relative phase encoding,) does not provide orthogonal channels. Thus, an interference-cancellation technique (e.g., phase-space beam forming) is required to enable phase-encoded channels. However, in systems that use redundantly modulated carriers (e.g., CI), phase encoding can be used to provide orthogonal or quasi-orthogonal phase-space channels. The operation of a correlator receiver may be illustrated mathematically by considering an application of a phase-space function $\Phi_n(f_i, \phi_j)$ to a received signal $R(t,f,\phi)$ in a matched filter. The correlation process is described by:

$$\left(\Phi_n\left(f_i,\phi_j\right)\right)\otimes R(t,f,\phi)=F_1+F_2+F_3$$

The term F_1, represents a recovered desired data stream $s_n(t)$ in the phase space described by $\Phi_n(f_i, \phi_j)$:

$$F_1=s_n(t)\sum_{\substack{i=i'\\ j=j'}}\left(f_i,\phi_j\right)\otimes\left(f_{i'},\phi_{j'}\right)$$

The summation is normalized to one. Thus $F_1 = s_n(\mathrm{t})$. The term F_2 represents interference between carriers having different frequencies f_i. Thus, for orthogonal carrier frequencies:

$$F_2=s_{n'}(t)\sum_{i\neq i'}\left(f_i,\phi_j\right)\otimes\left(f_{i'},\phi_{j'}\right)=0$$

The term F_3 represents interference resulting from carriers having the same frequencies f_i, but different phases ϕ_j. The term F_3 is described by:

$$F_3=s_{n'}(t)\sum_{\substack{i=i'\\ j\neq j'}}\left(f_i,\phi_j\right)\otimes\left(f_{i'},\phi_{j'}\right)$$

The present invention codes subchannels with a selection of phase spaces $\Phi_n(f_i, \phi_j)$ such that F_3 is substantially small or null (i.e., the subchannels are quasi orthogonal or orthogonal).

FIG. 25A through FIG. 25K show correlations between two orthogonal phase spaces $\Phi_n(f_i, \phi_j)$. Each of FIG. 25A to FIG. 25K shows a pair of same-frequency carriers having phase relationships defined by the phase spaces $\Phi_n(f_i, \phi_j)$. In each figure, the carrier frequencies f_i are different. In this case, a phase difference $\Delta\phi_j$ between each pair of carriers is defined by a translation of a pulse width:

$$\Delta\phi_j = \phi_{jo} + (1/(f_s N))2\pi f_j$$

where ϕ_{jo} is an initial phase (in this case $\phi_{jo} = 0$) and $1/(f_s N)$ is the duration of a pulse resulting from an in-phase superposition of the N carriers. A sum of the correlations in FIGs. 25A to 25K is substantially zero.

There is a large number of orthogonal phase spaces $\Phi_n(f_i, \phi_j)$ defined by phase differences $\Delta\phi_j$ and initial phases ϕ_{jo} that may be selected. Either or both the initial phase ϕ_{jo} and the set of applied phase shifts may be selected so as to reduce the peak magnitude of the carrier superposition. Phase spaces $\Phi_n(f_i, \phi_j)$ may be selected with respect to particular coding requirements.

Phase spaces $\Phi_n(f_i, \phi_j)$ may be provided with time dependence in either or both the carrier frequencies f_i and the carrier phases ϕ_j. For example, frequencies may be varied with respect to time to provide power control and/or interference mitigation. Phase differences $\Delta\phi_j$ and/or initial phases ϕ_{jo} may be provided with time dependence in order to enhance security, diversity, signal quality, and/or system capacity.

A second set of orthogonal phase spaces $\Phi'_n(f_i, \phi_j)$ is created by providing a phase offset of $\Delta\phi'_j = (1/(f_s N))\pi f_i$. Each set of orthogonal phase spaces is quasi-orthogonal with respect to each other. The correlations between quasi-orthogonal phase spaces are non-zero. Correlations may be regarded as diversity-parameter spaces when applying interference cancellation, multi-user detection, and beam-forming techniques. For example, weights may be applied to correlations and/or correlated signals to cancel overlapping correlations.

10. Array Processing with a CI Architecture

Embodiments of the invention may operate with a combination of CI processing with a plurality of transmitter elements with one receiver element, or a plurality of receiver elements with one transmitter element, or a plurality of both transmitter and receiver elements. CI processing decomposes a time-domain signal into a plurality of redundantly modulated frequency channels. Multiple data symbols and/or data streams redundantly modulate a common set of carrier frequencies. The signals received at a receiver in one frequency channel (or bin) of a CI process are substantially correlated with signals received in at least one other bin of the CI process.

One form of a CI process is a procedure that plays a part in transforming a time-domain sequence (e.g., a data stream) into a parallel set of substantially orthogonal channels (or bins) wherein the signals in one bin substantially interfere with the signals from other bins in a superposition process. In other embodiments of the invention, the channels may be quasi-orthogonal or non-orthogonal. Typically, the superposition process involves combining the interfering signals in a way that provides constructive interference (or preservation) of one or more desired data symbols or data streams and destructive interference (or cancellation) of one or more undesired or interfering signals. CI may include processes such as, but not limited to correlation, convolution, interferometry, and weighting.

A CI process converts a high-speed, serial data stream into a plurality of low-speed, parallel data streams. This conversion reduces the effects of intersymbol interference that occur in a multipath environment. Guard bands and cyclic prefixes are well known in the art and may be applied to CI architectures. The orthogonal channels each have relatively narrow bandwidth compared to the bandwidth of the serial data stream. Thus, array processing is simplified when a CI architecture is employed.

According to the present invention, CI may be used to create multiple independent spatial subchannels within one or more frequency channels or bins, even in the presence of significant levels of interference. Multiple communication channels may be deliberately overlaid in phase space and the interfering phase-space channels may be separated via spatial processing in each of the frequency channels. Thus, substantial improvements in system capacity may be achieved by spatial processing in the CI frequency channels. Spatial processing may be performed in a CI receiver to remove the effects of interference from undesired receivers and to enhance reception of desired signals.

Multipath and other propagation effects tend to be frequency dependent. Therefore, the effectiveness of beam forming will vary with respect to frequency, making some spatial subchannels more useful than others. Superposition averages the effects of subchannel performance across a CI phase space, thus avoiding the complexity of bit loading and other adaptive techniques required by OFDM. Since CI can provide the

fundamental signal structure for any conventional transmission protocol, the benefits of CI may be realized without substantial changes to legacy communication systems.

11. Transmitter System

FIG. 26 illustrates a transmitter in accordance with one embodiment of the invention. An input signal $s(t)$ including any form of digital and/or analog data is coupled into an encoder/interleaver system 10 that encodes the input signal $s(t)$ into a symbol stream. The symbol stream is typically a sequence of complex digitized values. Each symbol may be a one-dimensional or multi-dimensional value. An exemplary one-dimensional symbol set is a PAM constellation. A complex one-dimensional symbol has in-phase and quadrature components. Thus, a QAM constellation is defined as a set of one-dimensional symbols. A sequential grouping of QAM constellation members is regarded as a multi-dimensional symbol set.

The encoder/interleaver system 10 is an optional system that can improve the bit-error rate of the transmitted symbols by enhancing redundancy and/or diversity. For example, time diversity, spatial diversity, polarization diversity, or the like may be introduced into the system. Encoding techniques may include one or more well-known methods, such as convolutional encoding with bit mapping to symbols, trellis encoding, block coding, cyclic redundancy, Reed-Solomon coding, and Automatic Repeat Queing.

An interleaving operation may be used to distribute the transmitted symbols over multiple diversity-parameter values (e.g., subchannels, phase spaces, etc.). Since CI provides redundant modulation of carriers (e.g., carrier frequencies), interleaving may be performed in the time domain and/or in the phase-space domain without sacrificing bandwidth efficiency. Interleaving distributes the effects of interference so that symbols degraded by the same interference are not grouped closely together in the sequence recovered by the receiver. In some applications, it may be advantageous to perform bit loading and/or power optimization within phase spaces based upon measured communication quality of the phase spaces. Power optimization and/or interference mitigation may be performed by adjusting carrier frequencies to avoid multipath fading and/or interference.

A training-symbol injector 20 may optionally place predetermined training-symbol values into the transmitted symbol stream. The training symbols may be used by

a receiver to estimate communication-channel parameters and/or determine spatial gain patterns or interference ratios of overlaid communication channels. A channel estimate may be used to aid in demodulation and decoding of the received data sequence. Measured interference ratios may be used to facilitate interference cancellation and/or optimization techniques based on some performance measure of the receiver channel outputs. The training sequence may be injected with respect to one or more diversity parameters, such as time, phase space, spatial subchannels, frequencies, etc. Blind adaptive spatial processing techniques may be performed, thus eliminating the inclusion of the training-symbol injector 20.

The training-symbol injector 20 is shown coupled to a CI/spatial processor 30. The processor 30 converts the input symbol stream to one or more CI signals. When the CI transmission process is combined with a CI reception process, at least one set of parallel bins (e.g., frequency channels) is created in which information transmitted in each bin substantially interferes with information transmitted in a plurality of other bins. The interference (which optionally may be combined with one or more interference cancellation techniques) enables separation of multiple data streams that are each redundantly modulated onto the same set of bins.

CI may be performed along with various techniques for creating multiple parallel channels. An IFFT at the transmitter may be combined with an FFT at the receiver. A filter bank may be used at both the transmitter and the receiver. Various transform pairs may be utilized, such as Hilbert transform pairs, wavelet transform pairs, etc.

The processor 30 may also perform spatial processing (e.g., beam forming). Spatial processing typically involves multiplying one or more symbols in a given transform bin or frequency band with one or more spatial-processing weights. The weights may be adapted or otherwise optimized with respect to a measured or otherwise predetermined performance indicator.

One transmitter embodiment optimizes the spatial weights so that multiple subchannels in a given bin can be converted at a receiver into substantially independent received spatial subchannels. The symbols from one received subchannel substantially interfere with symbols from other subchannels to reinforce desired signals and suppress undesired signals modulated onto the same set of bins. Another embodiment optimizes

the transmitter spatial weights to improve the received power of one or more spatial subchannels within one or more bins. Transmitter spatial weights may be adjusted to improve the average power of several spatial subchannels within several bins. Spatial weights may be adjusted to improve power within one or more phase-space channels. A further embodiment optimizes the transmitter spatial weights within a plurality of bins to increase power delivered to a desired receiver (or within a desired channel) while reducing interference radiated to unintended receivers (or channels). In another embodiment, spatial processing includes multiplying each symbol with a transmitter weight that is fixed for all bins. In this case, the weights are optimized to increase the time-averaged and/or frequency-averaged power transmitted to one or more desired receiver channels and reduce time-averaged and/or frequency-averaged power to undesired receiver channels. Another embodiment involves routing each symbol to an antenna array without any transmitter-applied weighting. One or more filters may be used in the processor 30 to shape the transmitted RF signal spectrum.

A parallel set of digital time-domain signals are coupled from the processor 30 to a modulation and RF system 40. The modulation and RF system 40 may include a set of independent RF up-converter chains that convert the digital baseband signal to a modulated RF carrier frequency. Upconversion may be accomplished using digital-to-analog converters, an RF mixer, and a frequency synthesizer. Upconversion may be accomplished by modulating a pulse generator and filtering the harmonic content of the pulse-generator output, as described in U.S. Pat. No. 5,955,992. Any type of upconversion may be provided by the modulation and RF system 40.

A transmit antenna array 50 couples the transmitted signal into a communication channel. Antenna arrays may include one or more co-polarized radiating elements. The antenna elements may be provided with multiple differential polarizations including. Polarizations may be linear, circular, and/or elliptical. Linear polarized transmit elements may be characterized by either or both static and time-varying polarizations. Circular polarized and elliptical polarized elements may include constant and/or time-varying angular velocities. Angular velocities of circular and/or elliptical polarized signals transmitted by the array elements may be similar or differential.

Multiple polarizations, multiple polarization rates, multipath effects, spatial diversity (occurring at either or both the transmitter and the receiver), and/or differences in characteristics of other diversity-parameter values allows the invention to create multiple subchannels within each bin. The invention may use diversity to enhance signal quality and/or system capacity. These enhancements may be provided to signal phase spaces via separate processing of phase-space components (i.e., individual carrier signals) prior to combining the components. Weights may be applied to the phase-space components to facilitate separation of overlapping phase-space signals. Signals in overlapping phase spaces may be processed to separate desired signals and/or remove interference.

12. Receiver System

FIG. 27 illustrates a CI receiver according to one embodiment of the invention. RF signals from each of a plurality of elements of an antenna array 110 are downconverted to digital baseband by a demodulation and RF system 120. The system 120 may include RF signal processing, digital-undersampling, low-pass filtering, envelope-detection, and/or matched filtering systems to effect downconversion. The system 120 also digitizes the received signals. A timing and frequency synchronization system (not shown) may be used to recover timing of the transmitted signals. Any synchronization technique may be used herein.

In some embodiments of the invention, the digital baseband signal is coupled to an optional channel ID block 130 and a receiver CI-spatial processor 140. The channel ID block 130 provides channel estimation (i.e., analysis and estimation of channel characteristics). Channel estimates may include a matrix corresponding to each bin. The matrix may include complex values that relate proportions of signals received at a plurality of receivers with respect to signals transmitted from a plurality of transmitters. Channel estimates may include a matrix corresponding to each phase space wherein off-diagonal values represent interference or overlap in phase spaces. The channel ID block 130 may provide channel determination, such as from a set of measured ratios of co-channel interference or spatial gain distributions determined from measurements of predetermined transmitted training sequences. Channel estimates for each bin are provided to the receiver CI-spatial processor 140.

Some embodiments of the invention provide the receiver with beam-forming capabilities. The structured nature of frequency-domain fading and/or time-domain fading may be exploited to effect adaptive beam forming. A channel impulse response may be characterized by combining channel responses for a plurality of frequencies. A single narrow frequency may be used to estimate a channel if the channel response is averaged over time. Exploiting the structured nature of frequency-domain fading can reduce the number of bins in which training information is transmitted. Time-domain correlation of the channel can be exploited to reduce the frequency of training-symbol insertion.

It is well known that both time-domain and frequency-domain correlations may be exploited separately. However, the present invention allows for the transmission of training sequences in orthogonal or quasi-orthogonal phase spaces, thus enabling simultaneous exploitation of time-domain and frequency-domain correlations. Since phase-space training symbols can occupy the same frequencies and time intervals as data transmissions, these training symbols allow real-time control and evaluation of the communication channel concurrent with other transmissions. Although the Channel ID block 130 is shown as a separate functional block from the receiver CI-spatial processor 140 and the decoder and deinterleaver block 150, they may share the same elements.

One of the steps performed by the receiver CI-spatial processor 140 includes providing interference between bins of received CI signals. The interference step may be performed to generate one or more superposition signals that are processed thereafter. The interference step may be performed after one or more initial processing steps (such as a set of correlations that correlate signals in each bin with one or more corresponding reference signals). A process of combining uncertainties in a decision process such that the uncertainties cancel or average may be regarded as an interference step. Typically, an interference step may provide either or both reinforcement (constructive interference) of a desired signal (or determination of a desired signal) and suppression of one or more interfering signals (or uncertainties in measuring a desired signal).

Another step that may be performed by the receiver CI-spatial processor 140 includes de-interleaving and/or decoding signals encoded across a plurality of diversity-parameter values. For example, the processor 140 may compensate for coding that

spreads redundantly modulated bins across multiple time intervals. Similarly, the processor 140 may decode signals encoded with respect to one or more diversity parameters including, but not limited to phase spaces, directionality, time, spatial separation, and diversity subspaces (e.g., spatial subspace, polarization subspace, etc.).

One or more steps performed by the processor 140 may include spatial processing. In one set of embodiments, a receiver spatial-processing step combines received signals within each bin using one or more vector-weighted inner-product steps to form one or more one-dimensional received spatial subchannels within each bin. The receiver weight vectors may be chosen to optimize at least one measured performance indicator (e.g., signal power, signal-to-noise ratio, signal-to-interference ratio, signal to noise-plus-interference, etc.). Receiver weights may be chosen with respect to one or more predetermined or measured parameters, such as interference ratios, spatial gain distributions, and/or subspace-channel encoding.

In one embodiment, wherein the transmitter and receiver have knowledge of channel state information within each bin, both transmitter spatial weight vectors and receiver spatial weight vectors may be chosen to optimize performance for a set of substantially independent subchannels within each bin. In another embodiment wherein the transmitter does not have channel state information, the receiver performs spatial processing to create a plurality of substantially independent subchannels within each bin. In another embodiment, the receiver reduces the effects of interference radiated by unintentional transmitters and performs spatial processing to create a plurality of substantially independent subchannels within each bin. In one example, the receiver includes a beam-forming stage that suppresses unintentional interference and a spatial-processing stage that creates subchannels within each bin for intentional interference. In another example, the receiver includes a single spatial-processing stage to suppress unintentional interference and create subchannels. In another example, the receiver creates at least one empty subchannel with respect to intentional interference and provides suppression of unintentional interference via one or more subsequent processing steps and/or stages. The embodiments may include adaptive and/or deterministic weights to increase time-averaged and/or frequency-averaged received power from one or more

desired spatial subchannels and/or reduce time-averaged and/or frequency-averaged interference power.

In one set of embodiments, spatial processing is performed with respect to each bin and then signals from spatial subchannels are combined to create phase spaces that may overlap, but exist in different spatial subchannels. In another set of embodiments, each antenna element processes a set of phase spaces. Interfering phase spaces and/or phase-space components are then separated via spatial processing. In either case weights may be provided to at least one set of bins and phase spaces.

Certain embodiments may involve simply passing samples received in each bin to the decoder and interleaver block 150 without performing spatial processing. Many spatial-processing techniques (also referred to as "beam forming") are described herein. However, other combinations of transmitter spatial-weight optimization techniques and receiver spatial-weight optimization techniques can be constructed around the principle concept of spatial processing in combination with CI.

The outputs of the processor 140 are coupled into the decoder and interleaver block 150 that outputs at least one estimated data stream s(t). In one embodiment, the block 150 decodes a symbol sequence that was encoded and transmitted through a plurality of bins and/or phase spaces with one or more substantially independent subchannels. The decoder includes an appropriate receiver counterpart for the combination of encoders selected for the transmitter. A preferred embodiment includes a deinterleaver, a trellis decoder or convolutional bit-mapping decoder employing a scalar weighted Euclidian maximum-likelihood sequence detector followed by a Reed-Solomon decoder, followed by an ARQ system to correct Reed-Solomon codeword errors. In another embodiment, the block 150 decodes a sequence of multidimensional symbols or groups of adjacent one-dimensional symbols received in each bin. The symbol sequences may optionally be transmitted with weighting that optimizes some measure of average signal quality.

In an alternate embodiment, trellis-encoded symbols may be grouped and interleaved such that symbols transmitted from multiple antenna elements within a given bin form a vector from a multidimensional QAM-encoder output symbol, or a sequence of one-dimensional QAM-encoder output symbols having adjacent locations in the pre-

interleaved encoder output sequence. Thus, a maximum-likelihood vector decoder may be constructed from a channel-matrix estimate within each bin and/or phase space. A maximum-likelihood decoder can compute a weighted vector Euclidean metric given a transmitted vector symbol trellis-state table, and deinterleaved received vector and deinterleaved matrix channel estimates from each bin and/or phase space. In either embodiment, the encoder polynomial and symbol constellation set may be optimized to improve the bit-error-rate performance given the characteristics of the channel fading that occurs in each bin and/or phase space. For example, a code polynomial optimization search may be provided by evaluating the product of the two norms of the vector difference between the correct transmitter symbol vector and the error-symbol vector.

Transmitter embodiments of the present invention may be adapted for use with one or more receivers coupled to a communication channel via one receiver element. Receiver embodiments of the invention may be adapted for use with one or more transmitters coupled to a communication channel via one transmitter element.

13. Multipath Channels

Communication channels may be characterized by multipath wherein each path has an associated fade and/or propagation delay. Multipath may be created by reflections in the channel environment, dispersion (e.g., chromatic, polarization, mode, etc.), interfering diversity-parameter values (e.g., cross polarization, overlapping frequencies, overlapping correlations, overlapping directionalities, etc.), and spatial separation of receiver arrays and/or transmitter arrays.

Multipath effects may be exploited to separate interfering communication channels. A unique aspect of the present invention involves creating multipath effects by redundantly modulating a plurality of data streams (or data symbols) on a set of carrier signals that are distinguished from each other with respect to at least one diversity parameter. A set of carriers may be distinguished from each other if they each have at least one unique diversity-parameter value, such as frequency, mode, polarization, etc. A set of carriers may be distinguished from each other if the set has a unique combination of diversity-parameter values. One example of a combination of diversity-parameter values is a group of relative or absolute phases defining a phase space. Other combinations of diversity-parameter values may be employed.

One broad set of embodiments of the present invention provides redundant modulation across a plurality of diversity-parameter values that are each characterized by different channel effects, such as multipath, dispersion, etc. Multiple data streams and/or data symbols may be modulated onto the same set of carriers characterized by the aforementioned diversity-parameter values. Data-stream and/or data-symbol transmissions may be associated with different locations in a communication channel. The diversity-parameter space may be exploited to provide at least one of channel-distortion (e.g., multipath fading) mitigation and capacity enhancements.

Distortion compensation may be provided by averaging the channel effects over a plurality of carriers via redundant modulation. Distortion compensation may be achieved by a transmission system that exploits channel effects to counter signal distortion applied by the transmitter.

Capacity enhancements may be achieved by exploiting the channel-specific distortion of a plurality of redundantly modulated carriers in a plurality of unique communication channels. Weighting and combining signals received from a plurality of channels may be performed to separate interfering signals and/or suppress undesired interference. A communication channel may be characterized by a combination of transmission characteristics (e.g., coding, directionality, diversity-parameter gain patterns, etc.), propagation effects (e.g., multipath, dispersion, path loss, scatter, diffraction, etc.), and receiver response. Thus, transmitters, receivers, and/or communication channels may be adjusted to optimize signal reception. In the case where a combination of diversity-parameter values are used to assist the processing of received signals, channel effects may change relationships between the diversity-parameter values. This change in relationship may be exploited to separate interfering signals or suppress interference, particularly when relationships convey data, enable channel separation, and/or provide channel-state information.

FIG. 28A illustrates a wireless communication scenario wherein a CI/spatial processing transmission system 230 coupled to a plurality *N* of transmitter elements 231.1 to 231.N transmits information into a wireless communication channel. The information is received by a plurality *N'* of receiver elements 331.1 to 331.N' and coupled into a CI/spatial processing receiver system 330. Multipath signal propagation

can result in spatially selective fading, delay spread, frequency-selective fading, and time fading. Spatial fading occurs as various received waveforms from different propagation paths combine with constructive and destructive interference at different locations. A receiver array (such as receiver array 331) located within a spatially selective fading field can sample the spatial gain distribution of the received signals at various locations so that each receiver element (such as receiver elements 331.1 to 331.N') detects different signal strengths.

The effects of multipath combining depend on the frequency of the multipath signals. Thus, signals having different frequencies will have different signal strengths at a given location. Delay spread results from different propagation path lengths. If the relative distance between the transmitter elements and the receiver elements is changing (such as due to relative motion between the transmitter and receiver and/or motion of objects that influence multipath effects), then the multipath effects will change with respect to time. Time-varying multipath may be analyzed to adapt either or both transmitter array processing and receiver array processing.

Multipath also affects other diversity-parameter values. For example, signal phase varies with respect to delay spread. The phase response of a received signal in a multipath environment can vary with respect to spatial location and frequency. Relative phase between different signals can vary due to dispersion. Received polarization varies with respect to depolarization and vector-combining effects in a multipath channel. Elliptical polarization states can vary with respect to path length as well as vector.

Multipath channels enable increased capacity because multipath induces a rank greater than one in the matrix channel present in each diversity-parameter bin. This enables the formation of multiple subchannels within each bin. In some cases, multiple diversity parameters may be exploited to mitigate multipath effects and increase channel capacity. For example, a plurality of redundantly modulated carrier frequencies may be combined to compensate for deep fades. The resulting time-domain signal may be processed along with interfering time-domain signals to enable signal separation. Signal separation increases bandwidth efficiency and/or suppresses undesired interference. A substantial advantage of the present invention is its ability to efficiently and economically exploit CI and spatial processing to achieve multi-dimensional diversity benefits.

Spatial separation between the transmitter elements (such as elements 231.1 to 231.N) can provide channel-distortion effects similar to multipath effects, even when reflected components are non-existent or negligible. Path loss may provide variable received power with respect to spatial location. Similarly, if each of a plurality of information streams is transmitted by a plurality of antenna elements, constructive and destructive combining provides space-varying fields.

FIG. 28B shows a transmit CI/spatial processing system 230 and a receiver CI/spatial processing system 330 each including a beam-forming system 232 and 332, respectively. In this case, transmission elements 231.1 to 231.N receive a plurality of weighted values of each data symbol. Since each data symbol is coupled to each of the transmission elements 231.1 to 231.N, the transmission elements 231.1 to 231.N may be regarded as a plurality of co-located transmitters. Each beam pattern generated by the transmitter's beam former 232 may be regarded as a separate antenna, which may have a unique directionality. Receiver elements 331.1 to 331.N' receive different proportions of each transmitted data symbol. The receiver elements 331.1 to 331.N' may be regarded as a plurality of co-located elements. The outputs of the receiver's beam former 332 may each be regarded as a separate receiver element that each receives different ratios of interfering signals. The interfering signals are separated and/or removed by the receiver CI/spatial processor 330.

Due to the difference in propagation delay between the various multipath components, a symbol transmitted into a communication channel is received as a collection of delayed copies. Thus, delayed and scaled versions of a symbol can interfere with other symbols. For high transmission rates or large delay spreads, inter-symbol interference causes substantial degradation in received-signal quality. In OFDM, large delay spreads result in large guard intervals and thus, reduced bandwidth efficiency and/or signal to noise.

The present invention reduces wideband transmission protocols to parallel low-rate channels without changing the time-domain characteristics of wideband protocols . This enables spread-spectrum protocols to benefit from narrowband operations, such as beam forming. Each of the low-rate channels may be processed to provide a plurality of subspace channels to enhance system capacity. The present invention may use an

adaptive beam-forming front end or a multi-user detector to remove inter-symbol interference. Nulling inter-symbol interference can enhance bandwidth efficiency and signal quality.

FIG. 28C shows a transmitter array 231 and a receiver array 331 each having a plurality of polarizations. Each signal received by the receiving antennas 331.1 to 331.N' may have an arbitrary amount of cross-polarization interference. In a multipath environment (or an environment that has multipath effects), there is an additional increase in the channel matrix rank in each bin that can be exploited to further increase capacity and/or improve signal quality. In this case, the transmitter array 231 is shown with three ($N = 3$) elements 231.1, 231.2, and 231.N having polarizations in orthogonal directions. Similarly, the receiver array 331 is shown with three ($N' = 3$) elements 331.1, 331.2, and 331.N' having polarizations in orthogonal directions. However, the transmitter elements 231.1, 231.2, and 231.N and the receiver elements 331.1, 331.2, and 331.N' may be oriented in non-orthogonal polarization directions. Furthermore, either or both the number N of transmitter elements and the number N' of receiver elements may be greater than or less than three.

FIG. 28D shows a generalized communication system wherein a transmitter 230 generates a set of CI carrier signals modulated by a plurality N' of data-symbol streams $s_{n'}(t)$ wherein an n^{th} carrier is distinguished by at least one diversity-parameter value Λ_n. Each diversity parameter Λ_n is weighted by the transmitter with a weight $\alpha_{n'n}$ corresponding to an n'^{th} symbol stream $s_{n'}(t)$ and an n^{th} carrier. The applied weights provide an effective multipath channel that may be processed by a receiver to create a plurality of Λ_n-spatial subchannels. Thus, the invention provides an efficient method for exploiting the multi-dimensional nature of a multi-valued diversity-parameter channel.

The number of possible spatial subchannels may equal the number N of diversity-parameter values Λ_n. An exception to this rule occurs when one diversity parameter is coded with respect to a different diversity parameter. For example, it has been shown that multiple carrier frequencies may be combined into a number of time-domain signals that exceeds the number of carriers. Multi-channel detection in the time domain may provide a number of temporal spatial subchannels that exceeds the number of carrier frequencies.

A single antenna element 230 and 330 is shown for both the transmitter 231 and the receiver 331. The number of possible subchannels is related to the number N of diversity- parameter values Λ_n and the diversity parameter in which received-signal processing is performed. If a plurality of antenna elements (not shown) is employed at both the transmitter 230 and the receiver 330, then multipath effects can provide an increase in the channel matrix bin and thus, further improve channel capacity and/or signal quality.

A receiver may be instrumental in generating and optimizing diversity parameter spatial subchannels by adjusting its response to one or more diversity-parameter values. A plurality of receiver responses may be frequency dependent to facilitate the separation of interfering signals distributed across different frequency bands. Receivers may be tuned to different polarizations, modes, matched filters, or the like to optimize interference cancellation and suppression. Time-domain processing may be performed using a matched-filter technique wherein different reference signals correlated with a received signal produce different correlations. Overlapping signal correlations may be removed using a weight-and-sum technique or any other appropriate method of interference removal.

FIG. 28E shows a plurality of transmission systems 230.1 to 230.N' separated with respect to at least one diversity parameter. For example, the transmission systems may be spatially separated. Each transmitter 230.1 to 230.N' generates a plurality of redundantly modulated carrier signals. The carrier signals are distinguished by different values of at least one diversity parameter. A single receiver 330 is shown. However, a plurality of receivers (not shown) or a plurality of receiving elements (not shown) may be provided wherein the receivers or receiving elements are separated with respect to at least one diversity parameter.

In the case of spatially separated transmitters and/or receivers, signals may undergo multipath and/or multipath-like effects in the propagation channel, the transmitter systems, and receiver systems. The effects of multipath on the transmitted signals can be characterized by complex weights provided to the carrier signals. Transmitters of the present invention may provide complex weights to the carriers. A variable responsiveness of the receiver(s) with respect to one or more diversity

parameters may also provide complex weighting. A desirable effect of the complex weighting is to enable the creation of diversity-parameter subchannels that are specific to spatial locations and/or transmitters. A receiver demodulates or otherwise extracts data symbols from the carriers for each bin (i.e., diversity-parameter value). The demodulated data symbols may be weighted and combined or otherwise processed to separate and/or remove interfering symbols.

MIMO architectures are illustrated with respect to multiple transmitter elements and multiple receiver elements. MIMO architectures are also illustrated with a single transmitter element and a single receiver element wherein certain types of signal processing by either or both the transmitter and the receiver provide an effective alternative to using transceiver arrays. The MIMO techniques described herein may be adapted to multiple-input, single-output (MISO) architectures and single-input, multiple-output architectures.

14. CI-Space Processing

FIG. 29A shows a plurality of waveforms 211, 212, and 213 that may be transmitted by a first transmitter. The waveforms have a common frequency f_1. However, the waveforms have unique phases ϕ_{11}, ϕ_{21}, and ϕ_{31}. The waveforms 211, 212, and 213 may include guard intervals and/or cyclic prefixes to mitigate inter-symbol interference. A receiver (not shown) may use a matched filter to detect symbol values modulated onto the waveforms. However, a receiver tuned to the frequency and phase of a particular waveform will receive interfering contributions from any same-frequency waveform having a phase offset $\Delta\phi$ that is not equal to $\pi/2$.

Proportions of interference from received phase-offset, same-frequency waveforms are easily determined. In one exemplary case, known phase offsets between each of a plurality of received waveforms and a reference signal used in a matched filter enable a simple determination of interference ratios. The correlation of a reference signal with a same-frequency received signal is shown in FIG. 29B over a phase-offset interval of 2π. Other methods of determining interference ratios and measuring interference ratios may be used. For example, these methods may employ adaptive techniques. The important thing to note is that, although the absolute phases of the received signals may

vary due to delay spread in a multipath environment, the relative phases between same-frequency signals will remain unchanged.

FIG. 30 shows a communication system according to one embodiment of the invention. A signal generator 401 generates at least one carrier signal having at least one carrier frequency f_n. A phase-channelizing circuit 402 provides a plurality of unique phases (e.g., ϕ_{11}, ϕ_{12}, and ϕ_{13}) to a plurality of copies of the at least one carrier signal. The phase-channelizing circuit 402 may provide phase shifts and/or delays to the carrier-signal copies. The output of the phase-channelizing circuit 402 is coupled to a transmit module 403. The transmit module 403 may perform functions that are typically associated with transmitting an information signal into a communication channel. For example, the transmit module 403 may perform one or more of the following functions: upconversion, mixing, filtering, amplification, sub-carrier modulation, encoding, beam forming, etc.

A receiver 450 receives the transmitted signals. A receiving unit 451 couples the received signals out of the communication channel. The receiving unit 451 may perform conventional receiving functions, such as (but not limited to) amplification, mixing, downconversion, decoding, demodulation, filtering, beam forming, etc. The received signal is processed in a correlator bank 452. The received signal may be split into a plurality of signals and correlated with a plurality of reference signals. The reference signals are generated using a signal generator 453 that produces at least one initial signal having at least one frequency f_n. The initial signal is split or otherwise replicated into a plurality of reference signals that are provided with different phases in a phase-adjustment circuit 454. Each reference-signal phase may be locked onto at least one received-signal phase. Locking may be performed using adaptive techniques and may employ a feedback loop (not shown). Outputs from the correlator bank may include a plurality of interfering data symbols. A multi-channel detector 455 may separate the interfering data symbols.

FIG. 31A shows a plurality N of transmitters 400.1 to 400.N that are spatially separated in a multipath environment. Each of the transmitters 400.1 to 400.N transmits a plurality of signals redundantly modulated on at least one similar carrier frequency. The transmitters may each use an identical constellation of phases (e.g., ϕ_{11}, ϕ_{12}, and ϕ_{13}) for

at least one frequency f_n. The spatial separation of the transmitters 400.1 to 400.N ensures that transmitted signals from each transmitter experience unique propagation effects. This enables the interfering received signals to be separated, even though signals corresponding to each transmitter and frequency f_n are already interfering. Spatial processing at either or both the transmitter and the receiver enables interfering inter-transmitter carrier frequencies to be separated before intra-transmitter carrier signals are separated. Alternatively, the propagation effects of signals transmitted by multiple transmitters having unique diversity-parameter values may be processed concurrently with intra-transmitter carrier signal processing or adapted to intra-transmitter carrier signal processing methods.

A receiver 450 includes an optional beam former 456 coupled to an array of receiving elements 458.1 to 458.N'. Each of the elements may be processed independently using any well-known technique, such as interference cancellation. The elements may be processed together using conventional beam-forming techniques. In either case, signals transmitted from each of the transmitters 400.1 to 400.N have unique propagation characteristics that enable the creation of spatial subchannels. Complex weighting at either or both the transmitters and the receiver 450 may create these spatial subchannels. Each spatial subchannel can then be processed to separate intra-subchannel interference that results from the selected constellation of phases (e.g., ϕ_{11}, ϕ_{12}, and ϕ_{13}).

FIG. 31B shows a communication system in which a transmitter 400 includes a plurality N of transmitter elements 408.1 to 408.N. In this case, beam forming may be performed by the transmitter 400 to facilitate generation of a plurality of spatial subchannels.

15. CI Procedures

FIG. 32 shows a plurality N of carrier frequencies wherein each carrier frequency f_n (n=1,...,N) has a plurality of subcarriers. Each subcarrier is distinguished by one of a plurality M of phases ϕ_{mn} The subcarriers may be modulated and transmitted by a single transmitter. In this case, multipath effects (which cause frequency-dependent amplitude and phase variations to the carrier frequencies) do not affect the relative phases ϕ_{mn} and relative amplitudes of the group of subcarriers transmitted by the single transmitter.

Groups of subcarriers transmitted by other transmitters may experience transmitter-specific multipath effects (such as due to corresponding propagation environments, transmitter-impressed effective multipath and/or coding, and/or differential receiver responses to signals transmitted by different transmitters). Transmitter-specific multipath effects may be exploited by the invention to separate interfering groups of subcarriers. For example, spatial processing (or any other type of appropriate diversity-parameter processing may be performed to separate signals transmitted by a plurality of transmitters.

In a preferred embodiment of the invention, the carrier frequencies are orthogonal. Thus, the carrier frequencies are separable via conventional signal-separation techniques, such as filtering and correlation. In another preferred embodiment of the invention, the carrier frequencies are non-orthogonal or quasi-orthogonal. Thus, data symbols modulated onto different non-orthogonal carrier frequencies may be separable using multi-user or multi-channel detection between two or more carrier frequencies.

An embodiment of the invention also enables separation of a plurality of data symbols modulated onto the interfering subcarriers corresponding to each carrier frequency f_n. For example, matched filters may be used to provide differing proportions of interfering signals to a multi-channel detector 455 (e.g., a weight-and-sum canceller, matrix diagonalizer, or the like). In this case, the different proportions correspond to different carrier phases ϕ_{mn}. Other diversity parameters (as well as diversity parameter combinations) may be utilized within a carrier frequency.

The multi-channel detector 455 separates the interfering signals and thus, enables a system capacity that approaches N•M. This corresponds to a bandwidth efficiency of M symbols/Hz. Spatial processing further enhances system capacity. If K spatial subchannels are created, system capacity may be increased by a factor of K. If an L-ary differential-modulation scheme is employed, system capacity and bandwidth efficiency may be multiplied by the value L. Maximum system capacity C_s is expressed by:

$$C_s = N \cdot M \cdot K \cdot L$$

and maximum bandwidth efficiency e_{BW} is expressed by:

$$e_{BW} = M \cdot K \cdot L\,.$$

Diversity may be exploited to provide capacity or signal-quality benefits. In some cases (such as in interference-limited spread-spectrum systems), signal quality and system capacity can be improved simultaneously. Since multipath fading may cause certain frequencies and/or spatial locations to be unusable, the maximum system capacity C_s and bandwidth efficiency e_{BW} may not always be attainable. Redundancy may be employed in one or more diversity spaces to optimize the bit error rate. For example, multiple antennas may be used by the transmitter and/or the receiver to provide spatial redundancy. Similarly, different carrier frequencies may be modulated with the same data symbols.

In one embodiment, the carrier signals shown in FIG. 32 may be redundantly modulated. The different-frequency carriers corresponding to phases ϕ_{11}, ϕ_{12}, and ϕ_{13} may be modulated with a first data symbol. Phases ϕ_{21}, ϕ_{22}, and ϕ_{23} may be modulated with a second data symbol. Phases ϕ_{31}, ϕ_{32}, and ϕ_{33} may be modulated with a third data symbol. This redundancy allows a receiver to select a preferred result from a set of three processes. Thus, fading or interference that substantially degrades one or two of the carrier frequencies does not substantially affect the bit-error rate of the received symbols. In one embodiment, the effects of interference and signal distortion at one or more carrier frequencies are distributed evenly to all users such that no user experiences substantial debilitation of their channel. In another embodiment, carrier frequencies may be changed to avoid deep fades and/or interference. Carrier-frequency adjustment may be performed as a means of power control.

An alternative method for processing the received signals includes correlating the phase-frequency combinations for each data symbol and summing the correlation signals. A data symbol may be modulated (or otherwise impressed) onto a single carrier phase ϕ_{mn} corresponding to an m^{th} phase in an n^{th} frequency. A k^{th} data symbol may be modulated onto a k^{th} set of carrier phases $(\phi_{mn})_k$ corresponding to multiple m phases and multiple n frequencies. One or more of the correlation signals may be weighted with a complex weight, a vector of complex weights, or a complex-weight matrix prior to summing.

Each subcarrier phase within a carrier frequency may experience interference from at least one other subcarrier phase. In one case, a k^{th} data symbol (or data channel) is redundantly modulated onto multiple carrier frequencies with respect to a k^{th} group of

phase shifts $(\phi_{mn})_k$. Subsequent data symbols are redundantly modulated onto the same set of frequencies. For example, a k'^{th} data symbol (or data channel) is redundantly modulated onto the same carrier frequencies with respect to a different (k'^{th} group) of phase shifts $(\phi_{mn})_{k'}$. The total interference $(I_{k'k})_T$ between the k^{th} data symbol and the k'^{th} data symbol is expressed by:

$$(I_{k'k})_T = \sum_n (I_{k'k})_n$$

where interference between the data symbols is summed over the n carrier frequencies.

Since the interference $(I_{k'k})_n$ depends on the relative phase difference $\Delta\phi_{mn} = \phi_{m'n} - \phi_{mn}$ between the m^{th} and m'^{th} phases for an n^{th} carrier frequency, the phases ϕ_{mn} may be selected such that the sum of the interferences is zero or very small:

$$(I_{k'k})_T = \sum_n (I_{k'k})_n = 0$$

In another embodiment of the invention, the interference terms may be weighted with complex weight values w_{mn} such that the sum of the interferences is zero or very small:

$$(I_{k'k})_T = \sum_n w_{mn}(I_{k'k})_n = 0$$

Each complex weight value w_{mn} may include a single complex values, a vector of complex weights, or a matrix (or higher-order distribution) of complex weights.

In FIG. 32, a superposition signal S_k corresponds to a k^{th} group of phases $(\phi_{mn})_k$ for a special case in which $m=k$ and the relative phases ϕ_{mn} within each n carrier are associated with incremental translations (in time) of the carrier. Thus, phases ϕ_{k1}, ϕ_{k2}, ..., ϕ_{kN} are redundantly modulated with a k^{th} symbol $s_k(t)$. In this case, there are three superposition signals S_1, S_2, and S_3 that are centered at times t_1, t_2, and t_3, respectively. The phases ϕ_{1n} of the carrier frequencies f_n that correspond to the first symbol $s_1(t)$ provide a maximum for each n carrier at time t_1. Similarly, the second symbol's $s_2(t)$ phase relationship ϕ_{2n} provides for in-phase carriers at time t_2. The third phase relationship ϕ_{3n} provides in-phase constructive combining between carriers at time t_3. In this case, the superposition signals S_1, and S_3 do not overlap and, therefore, are orthogonal in the time domain. Consequently, the phase relationships ϕ_{km} of the

corresponding component carriers ensure orthogonality between the redundantly modulated symbols $s_k(t)$.

The relative phase of each carrier corresponding to a k^{th} orthogonal superposition signal S_k is related to an integer multiple of the superposition-signal width:

$$\phi_{mn} = \left(\frac{1}{f_s N} \right) 2\pi f_n m$$

where $(1/f_s N)$ expresses the width. The term f_s is the frequency separation and N is the number of carrier frequencies. An n^{th} carrier frequency is expressed by f_n.

The orthogonality of superposition signals (such as the superposition signals S_1 and S_3) demonstrates one class of orthogonality requirements for phase relationships ϕ_{mn}. However, the present invention may use orthogonal phase relationships that result in superposition signals that are not orthogonal in the time domain. For example, phase offsets may be provided to each carrier in order to reduce the peak-to-average power of the superposition signals. Furthermore, various carrier-frequency and/or carrier-amplitude distributions may be used that support orthogonal or quasi-orthogonal phase spaces, but may not provide orthogonality in the time domain.

The use of quasi-orthogonal coding is well known in spread-spectrum techniques and is equally applicable to CI. FIG. 32 also illustrates one form of quasi-orthogonal phase coding that is illustrated by quasi-orthogonal time-domain superpositions S_1, S_2, and S_3 of the carrier signals. The relative phase of each carrier corresponding to a k^{th} superposition signal S_k is related to an integer multiple of half the superposition-signal width:

$$\phi_{mn} = \left(\frac{1}{f_s N} \right) \pi f_n m$$

Quasi-orthogonal phase relationships employed by the present invention may not have a quasi-orthogonal corollary in the time domain. For example, phase offsets may be provided to each carrier in order to reduce the peak-to-average power of the superposition signals. Various carrier-frequency and/or carrier-amplitude distributions may be used that support orthogonal or quasi-orthogonal phase spaces, but may not enable quasi-orthogonality in the time domain.

Several embodiments of the present invention use space-frequency processing at either or both the transmitter and the receiver. This processing substantially eliminates inter-symbol interference caused by channel correlations across space (antenna correlation) and time (delay spread). This processing may include one form of CI processing, which separates a wide-bandwidth, high data rate communication signal into a plurality of small-bandwidth, low data rate channels without changing the time-domain characteristics of the wide-bandwidth signal. For example, DS-CDMA may be decomposed into component signals at a transmitter and processed using advantageous parallel-processing and/or narrowband-processing techniques. A CI DS-CDMA signal may be received by a receiver and processed in any of several ways. The received CI DS-CDMA signal may be processed as a conventional DS-CDMA signal. The received signal may be processed with respect to its narrowband components in an array-processing operation or the like. A CI receiver may be used to process CI DS-CDMA components and exploit capacity and signal-quality benefits of CI.

FIG. 33A shows an arrangement of N incrementally spaced-in-frequency carrier signals. This case illustrates one set of N carriers capable of having up to N orthogonal phase spaces wherein each of the phase spaces spans the entire set of carriers. The symbol duration T_s is equal to the inverse of the carrier-frequency spacing f_s. In a first embodiment of a CI architecture, a single user may use all of the phase spaces. In a second embodiment, a plurality of users may use the same set of carriers but be assigned to different phase spaces.

FIG. 33B shows an arrangement of N incrementally spaced-in-frequency carrier signals divided into M sets of carrier signals. In this case, each of the M carrier sets includes N/M carriers. There are N/M orthogonal phase spaces per set of carriers. The symbol duration T_s is equal to the inverse of the carrier-frequency spacing f_s. Carrier-frequency spacing in each carrier set may be selected to reduce correlated fading between adjacent carrier frequencies.

In a third embodiment of a CI architecture, each user is assigned to one of the M carrier sets. In one case of this embodiment, multiple users may be assigned to the same carrier set. Each user sharing a particular carrier set may be assigned to one time slot and/or phase space. Alternatively, each user sharing a carrier set may use more than on

time slot and/or phase space. In another case of the third embodiment, only one user is assigned to each carrier set and that user can use one or more of the available phase spaces and/or time slots.

In a fourth embodiment of a CI architecture, each user is assigned to a plurality of the M carrier sets. In one case of the fourth embodiment, each user is assigned to the same constructive-interference time slot (or equivalent phase space) for each of the carrier sets to which the user is assigned. In another case of the fourth embodiment, each user is assigned to a different time slot and/or phase space relative to each carrier set. In another case, each user is assigned to a plurality of time slots and/or phase spaces for each carrier set. This plurality of time slots and/or phase spaces may include similar or different time slots and/or phase spaces relative to the different carrier sets.

In a fifth embodiment of a CI architecture, multiple users are assigned to the same time slot(s) and/or phase space(s) relative to each of one or more carrier sets. In this case, a code is applied to each phase space and/or superposition signal to differentiate users. For example, code chips applied to superposition pulses can provide a DS-CDMA signal in the time domain. In another example, the application of code chips to frequencies that comprise a phase space can enable information symbols modulated in that phase space to be separable from other symbols modulated in that phase space with a different code.

One advantage of the class of CI architectures illustrated in FIG. 33B is that it allows a multicarrier protocol consisting of N carriers to be separated into smaller groups of carrier signals, which are easier to process. The smaller groups also provide substantially the same frequency-diversity benefits as the full N-carrier architecture. The size of these groups and the carrier frequencies utilized by each group may be adapted with respect to system and environment factors, such as fading, distortion, and interference.

FIG. 34A shows a single carrier frequency f_n having a set of quasi-orthogonal phases $\phi_{(mn)k}$ used in a quasi-orthogonal phase-space division multiplexing technique. In this case, a k^{th} user is provided with M phase channels. Alternatively, the phase channels may correspond to different users. The advantage of assigning all of the phase channels for a given frequency to a single transmitter is that each of the phase channels undergoes the same multipath effects and interference in a communication channel. Known

correlations between the phase channels are used to help separate interfering symbols transmitted on those channels. However, interference and multipath distortion do not substantially affect correlations between the phase channels.

FIG. 34B shows a plurality of carrier frequencies f_n to $f_{n'}$ having sets of quasi-orthogonal phases $\phi_{(mn)}$ to $\phi_{(mn')}$. In this case, the number M of phase spaces is greater than the number N of carrier frequencies. Multiple carriers may be used to increase capacity or mitigate the effects of fading and interference via redundant modulation. In one case, the phases of each carrier may be selected with respect to incremental time offsets. In another case, the carrier phases may be incrementally spaced in phase. For example, binary phase-shift key modulation may be implemented with carrier phases spaced throughout an interval of 0 to π. Quadrature shift-key modulation may be implemented with carrier-phase channels throughout an interval of 0 to $\pi/2$. Differential amplitude modulation may also be performed in carrier-phases channels.

Coding may be provided that distributes a data symbol over a plurality of carrier frequencies, time intervals, and/or phases. Such coding techniques are commonly employed in OFDM. Similarly, guard bands, cyclic prefixes, and various signal-processing techniques used with OFDM and other multicarrier protocols may be provided to CI signals.

For $M > N$, multi-user detection is typically used to separate interfering data symbols. A multi-user detector may sample multiple phase offsets ϕ_{mn} for each carrier frequency. Sampling may be achieved via matched filtering. For redundantly modulated carriers (i.e., multiple frequency channels modulated with the same data symbol), either or both phase-domain sampling and time-domain sampling may be used. Time-domain sampling may involve combining the modulated carriers to generate superposition signals that can be separated via time-domain multi-user detection. Multi-user detection may include interference cancellation, interference suppression, constellation processing, and/or optimization.

Another embodiment of CI is characterized by a mapping process performed by a receiver that maps a time-domain sample of a high-bandwidth signal into a parallel set of low-bandwidth, long-duration symbols. For example, a serially transmitted, high bandwidth signal shown in FIG. 35A sampled over a sample interval t_s has a given

number P of data symbols u_p. FIG. 35B shows a frequency-domain plot of the sampled high-bandwidth signal shown in FIG. 35A.

The time-domain sample of the high-bandwidth signal may be decomposed into a plurality N of carrier signals having different frequencies, as shown in FIG. 35D. Each carrier signal appears to be modulated with a complex data symbol v_n that is actually a complex-weighted superposition of symbols u_p. The carriers may or may not be incrementally spaced in frequency. The modulated carriers may or may not overlap. An n^{th} symbol v_n corresponding to an n^{th} carrier frequency f_n results from a superposition of complex weighted w_{np} data symbols u_p:

$$v_n = \Sigma_{p=1} w_{np} u_p$$

The vector representations v_n may be combined and processed using matrix operations to separate the data symbols u_p. Appropriate techniques involving any combination of estimation, optimization, and/or cancellation may be used to separate the data symbols u_p.

FIG. 35C shows a representation of N modulated carrier components. Each carrier frequency f_n is associated with a complex-valued symbol v_n. In this case, the frequencies f_n are incrementally spaced with a frequency separation of f_s between adjacent carriers. Each symbol v_n has a symbol period of $T_s = 1/f_s$. A corresponding frequency spectrum in FIG. 35D shows the N frequency components.

FIG. 35E shows a representation of N' modulated carrier components. In this case, $N' > N$. Thus, the frequency separation f'_s of the carriers is smaller than f_s. Consequently, the symbol v_n period T'_s is longer than T_s. The frequency separation f'_s and the symbol period T'_s are expressed by the following equations:

$$f'_s = f_s N/N'$$
$$T'_s = T_s N'/N$$

A frequency-domain spectrum of the carriers shown in FIG. 35E is shown in FIG. 35F.

One aspect of the invention includes selecting frequency components in order to optimize separation of the transmitted data symbols. Factors that may play a part in selecting carrier frequencies in a decomposition process include number of parallel operations, duration and frequency of sampling, and transformation characteristics between transmitted symbols u_p and received carrier symbols v_n.

FIG. 36 illustrates a CI process that may be used to process various types of wideband signals by decomposing the signal into discreet components and separating transmitted data symbols u_p embodied in values v_n associated with each carrier. Unlike a substantially orthogonalizing process (e.g., OFDM), a CI process converts a serial stream of data symbols into parallel channels (which are defined by different values of at least one diversity parameter) in which the transmitted data symbols interfere with each other. Each transmitted data symbols may be spread across multiple channels. Thus, one embodiment of a CI reception process includes an application of spatial processing in the frequency domain to separate transmitted data symbols. This spatial-processing application is referred to as frequency-diversity Interferometry in several of applicant's co-pending patent applications that are incorporated by reference.

In FIG. 36, transmitted data symbols u_p of a wideband signal are sampled in a sampling process 601. Sampling is performed over at least one time interval. A decomposition process 602 separates the sampled wideband signal into multiple component signals. Values v_n corresponding to each component signal are measured. A mapping process 603 is performed prior to sampling 601 and/or after decomposition and measuring 602. The mapping process 603 is used to determine vector relationships between the measured signals v_n and the transmitted data symbols u_p. The vector relationships are used to separate transmitted data symbols u_p in a separation process 604. The separation process 604 may include cancellation, constellation, estimation, and/or optimization processes to separate or otherwise determine the values of the transmitted data symbols u_p.

FIG. 37A illustrates a CI-TDMA signal. Time-shifted and scaled versions of a basic CI pulse shown in FIG. 37B may be used to generate time-domain signals that correspond to conventional TDMA protocols. Multiple closely spaced or overlapping pulses may be combined to generate longer-duration signals, as shown in FIG. 37C. Alternatively, each carrier may be provided with a complex weight to generate a particular time-domain signal.

CI-TDMA users may have different sets of frequencies. Thus, some users may share similar time slots and different carrier-frequency sets. Users that share the same carrier frequencies may be provided with different orthogonal or quasi-orthogonal time

slots. Differential amplitude and/or differential phase modulation may be employed with CI-TDMA.

FIG. 38 shows a CI-OFDM signal architecture for a plurality M of users. Each user is provided with a unique set of carriers. A user may be capable of using all available time slots/phase spaces associated with the set of carriers. Different users typically do not share the same carrier sets. However, coding may be provided that distributes a data symbol over a plurality of carrier sets. Other signal-processing techniques associated with OFDM may be incorporated into CI processing. A receiver typically employs at least one matched filter, which may be applied to individual carrier signals or to carrier superpositions. A receiver may provide time-domain processing, such as sampling, to received signals. For example, superposition signals may be sampled with respect to predetermined time references and time intervals.

Another form of CI-OFDM assigns a plurality of frequency-diverse carriers to serve as each OFDM carrier. This form ensures that a single carrier is not significantly affected by deep fades and narrowband interference. A further adaptation of this CI-OFDM form allows multiple signals to share each OFDM carrier.

FIG. 39 shows one of many possible CI MC-CDMA architectures. In this case, a k^{th} user having a k^{th} phase space ϕ_{kn} is provided with a spreading sequence c_{kn} in which a binary value is assigned to each of a plurality N of carriers. Multi-user detection may be employed to separate interference between the users.

FIG. 40A is a time-domain representation of a plurality K of DS-CDMA codes generated from a CI architecture. Each chip of the CI DS-CDMA codes is generated from a CI superposition of carriers. Each chip is provided with a binary code value corresponding to a DS-CDMA spreading code. The chip shape can be adjusted by applying weights to the chip's carriers. In this case, the chips have a distinctive sinc-like shape that allows chip overlap (shown in FIG. 40B). Chip overlap doubles the code length without increasing bandwidth. This can result in doubled capacity. Furthermore, CI provides enhanced frequency diversity, narrowband processing capabilities (e.g., spatial processing), and simplified signal processing at both the transmit and the receive sides.

FIG. 40C shows one of many possible carrier architectures for a CI DS-CDMA system. A k^{th} user is assigned a plurality M of phase spaces. Each phase space corresponds to a superposition of carriers that provide a chip for the k^{th} user's DS-CDMA code, which has a code length of M. Other users may share the same carriers and phase spaces. However, differences in users' codes provide for separation of the users.

FIG. 41 shows a CI architecture for Frequency-Hopping Spread Spectrum (FHSS). A k^{th} user is provided with a set of N time-dependent carrier frequencies:

$$f_k(t) = \sum_{n=1}^{N} f_{nk}(t)$$

Symbols are modulated onto a plurality of the carrier frequencies, which may change with respect to a time-dependent code. CI-FHSS symbols may be modulated over more than one time interval so as to create a time-diverse transmission signal. In one embodiment, received carriers may be processed separately and combined. Matched filters may be used to detect values on individual carriers. In another embodiment, a superposition of the carriers is processed to determine symbols modulated onto the carriers. In one case, a matched filter generates a superposition reference from time-dependent carriers.

Other spread-spectrum protocols may be created from a CI signal architecture. CI pulse-position modulation and other CI time-offset multiplexing techniques may be employed. The use of CI as an underlying architecture in other transmission protocols is an anticipated expression of the examples described herein. CI may be applied to hybrid protocols (e.g., multi-tone CDMA, CDMA-OFDM, synchronized CDMA, time-hopping direct sequence, time-frequency hopping spread spectrum, and frequency-hopped direct-sequence modulation), other spread-spectrum protocols (e.g., chirped spread spectrum, coherence multiplexing, wavelength division multiplexing, and discreet multi-tone), and non-spread spectrum protocols (e.g., frequency division multiplexing (FDM), including amplitude-modulated FDM).

16. Spatial Processing

Since a CI process can establish a plurality N of MIMO channels that are substantially independent from each other, spatial processing may be performed within each channel (or bin). FIG. 42 shows $K{\cdot}M$ symbols $s_{km}(t)$ modulated onto same-frequency

carrier signals. A plurality K of channels or users each modulate a plurality M of symbols onto the carriers relative to a plurality M of phases ϕ_{kmn}. The phases ϕ_{kmn} are identified relative to a k^{th} user or channel, an m^{th} phase corresponding the user and carrier, and an n^{th} carrier frequency f_n. The phases ϕ_{kmn} may or may not be unique with respect to user and/or carrier. The value m may be equal to or greater than one.

FIG. 43A shows K sets of M symbols $s_{km}(t)$ ($k = 1,\ldots,K$ and $m = 1,\ldots,M$) coupled into a plurality K of transmitter spatial weight (TSW) systems 310.1 to 310.K. These K sets of M symbols occupy frequency bin 1. Each member of a set of symbols $s_{km}(t)$ includes a data symbol and an associated subcarrier phase ϕ_{kmn}. Each TSW 310.1 to 310.K applies a weight vector to the set of symbols received at its input. In one configuration (shown in FIG. 43B), the TSW's apply weights to the sets of symbols before the symbols are modulated onto the carriers. In another configuration (shown in FIG. 43C), information-modulated carriers are weighted. The TSWs may be regarded as multipliers that multiply each set of symbols with a vector corresponding to a spatial direction in K-space. The K vectors define a subspace in K-space. The elements of the resulting vector are coupled to K summing junctions (or combiners) 311.1 to 311.K. The TSW vectors may be combined as column vectors to provide an orthogonalizing matrix or beneficial weighting matrix for the associated bin (e.g., frequency 1). Thus, for each bin, a vector including sets of M symbols $s_{km}(t)$ allocated to different spatial subchannels is multiplied by the orthogonalizing matrix to provide a resultant vector. Each element of the resultant vector includes sets of weighted symbols, which are transmitted by a corresponding antenna element. The weighted symbols are processed by a plurality of transmit processors 312.1 to 312.K. The transmit processors 312.1 to 312.K may provide well-known signal processing, including, but not limited to upconversion, mixing, filtering, encoding, amplification, automatic gain control, beam forming, etc.

FIG. 44A shows a CI-spatial receiver. A receiver spatial processor 340 includes a plurality K' of receiver spatial weight (RSW) systems 320.1 to 320.K' that each receive K' inputs, the inputs corresponding to a number K' of receiver 304.1 to 304.K' antenna paths. Within each RSW, a weight vector is applied to the inputs (e.g., an inner product is performed), thus producing a received signal sample $s'_{km}(t)$. The received signal samples include information symbols that are each associated with a specific subcarrier phase

ϕ_{kmn}. Since each sample $s'_{km}(t)$ may include a plurality of interfering signals, the samples $s'_{km}(t)$ are processed in a multi-user detector 244 to separate the interfering signals. The multi-user detector 244 may provide the receive side of a CI process by combining samples corresponding to different frequency bins to separate interfering signals, combining the processing results (e.g., correlations) of samples corresponding to different bins, and/or separating the sets of signals within each sample $s'_{km}(t)$ using weight and sum, cancellation, or any other appropriate interference-cancellation and/or interference-suppression technique.

An RSW vector has an associated direction in K'-space. Each RSW may be considered to be a multiplier in which the RSW vectors are row vectors. When the K' RSW vectors are collected into a matrix, the result is an output orthogonalizing matrix or beneficial weighting matrix for that bin. A vector that includes sets of symbols in a particular output bin is multiplied by the output orthogonalizing matrix to provide a vector including sets of symbols corresponding to different spatial directions. An appropriate selection of weights for either or both the TSWs and RSWs provides orthogonality to a number of spatial directions. If $K' \geq K$, the number of orthogonal spatial directions may be equal to K. Thus, the received sets of signals $s'_{km}(t)$ are substantially influenced only by corresponding transmitted signal sets $s_{km}(t)$.

FIG. 44B shows one of many possible embodiments of the multi-user detector 244. In this case, it is assumed that the carriers are redundantly modulated with information symbols. A superposition of the received sets of signals $s'_{km}(t)$ enables orthogonal or quasi-orthogonal distinction between redundantly modulated symbols encoded with respect to subcarrier phases ϕ_{kmn}. Optionally, complex weights w_k may be applied to the received signals to restore the transmitted relative phase relationships between the carrier frequencies, compensate for changes in signal magnitude that may occur during propagation and/or reception, and/or taper the received signals in order to reduce sidelobes. The weights may be single-value, vector, or matrix weights. The weights w_k may be adapted with respect to a feedback network or control signal. For example, measurements of the superposition signal may be provided such that phases of the weights are adjusted to optimize constructive interference.

Weighted signals are combined in a summing network 246 to produce superposition signals indicative of each of the modulated information symbols. The superposition signals are separated in a decision/detection system 247. Detection may include well-known reception techniques, such as, but not limited to sampling superposition signals within one or more predetermined time intervals, correlating the superposition signals with one or more reference signals over a predetermined interval, and performing difference measurements between predetermined time intervals.

FIG. 44C shows a matched-filter embodiment of the decision/detector system 247. Received superposition signals are multiplied by with at least one reference signal R_m. The product is integrated (or summed) over a predetermined time interval, such as a symbol duration T_s. The processed signal may be sampled. Optionally, the output of the matched filter may be compared to a constellation of values or some other reference (e.g., other output signals) to determine its value.

FIG. 44D shows another embodiment of the multi-user detector 244 for the case of redundantly modulated carriers. Matched filtering is used to process the received sets of signals $s'_{km}(t)$. For example, the received signals $s'_{km}(t)$ are combined with receiver-generated reference signals R_{km} to provide a plurality of correlation signals that include various interfering signals. The correlation signals are combined 246 and then processed via appropriate decision/detection methods 247. The received signals $s'_{km}(t)$ and/or the reference signals R_{km} may be weighted with complex weights to reduce interference.

FIG. 44E shows an embodiment of a section of a multi-user detector 244 that supports phase-encoded, non-redundantly modulated carrier frequencies. One of the received sets of signals $s'_{km}(t)$ $\{s'_{k1}(t)$ to $s'_{kM}(t)\}$ includes M signals corresponding to M subcarrier phases ϕ_{kmn}. The signal set $s'_{km}(t)$ is processed by M matched filters 245.1 to 245.M. Each matched filter 245.1 to 245.M produces an output vector having M values corresponding to different proportions of the signals $s'_{k1}(t)$ to $s'_{kM}(t)$. Weights w_{km} $\{w_{k1}$ to $w_{kM}\}$ applied to each matched-filter output vector enable separation of the interfering signals in a summing network 246. Each of the weights w_{k1} to w_{kM} may be expressed as a weight vector. Although a particular multi-dimensional weight-and-sum method is implied by FIG. 44E, other methods may be used to separate interfering signals. Even

beam-forming operations may be utilized in phase space to reduce interference. Feedback loops and optimization processes may be utilized to adjust one or more weights w_{km}.

Spatial processing may be applied to all *N* frequency bins to provide substantially independent spatial subchannels for each bin. Each spatial subchannel has a strength that may vary with respect to multipath and other conditions. The subchannels will have varying signal-to-noise ratios and thus, varying information-carrying capacity. Techniques for compensating for different subchannel strengths include bit loading and coding that involves redundant modulation. However, unlike redundant modulation techniques associated with CI, convolutional coding and error-correction coding (as well as other conventional redundant-modulation schemes) enable signal-quality enhancements by sacrificing bandwidth efficiency. Transmit power may also be adjusted to compensate for weaker subchannels.

Redundantly modulated CI signals provide frequency diversity benefits that enhance signal quality and increase bandwidth efficiency. Subcarrier power control can be avoided if the carrier signals provide enough frequency diversity that frequency-dependent fades do not substantially affect received data symbols. Alternatively, carrier frequencies may be adjusted, hopped, or simply avoided to reduce the effects of deep fades and narrowband interference. Non-redundantly modulated phase-encoded signals may be provided with conventional power control and subchannel-compensation techniques. Transitions between non-redundant and redundant signaling protocols may be employed.

Spatial processing techniques described herein have demonstrated cooperation between a transmitter and a receiver to effectively orthogonalize the spatial subchannels for each frequency bin. Alternatively, this orthogonalization may be provided by operations at only one end of the communication link. It may be advantageous to provide spatial processing at only one end of the link in order to reduce computational complexities at the other end(s). For example, only TSW system weights or RSW system weights are applied as spatial weights. Spatial processing at only one end of the link may be advantageous in cases where a channel model is used with spatial processing and the channel model is known only at that end.

Methods for determining weights based on a channel model include (but are not limited to) zero-forcing solutions and minimum-mean-square-error solutions. Many other solutions and variations exist that may be utilized in the present invention.

A transmitter can orthogonalize spatial subchannels to a given set of subscribers (e.g., distant receivers or a distant receiver array) if the transmitter has knowledge of the RSW system weights used to spatially process received signals from the subscribers. In a time-duplexed system, cancellation-duplexed system, and some other types of systems employing non-frequency duplexing, the channel exhibits reciprocity. Thus, TSW system weights can be learned from the RSW system weights when the RSW subspace directions are equal to the TSW subspace directions. For frequency duplexing, as with any non-reciprocal duplexing system, weights may be derived from channel estimations in which multiple instantaneous channel responses are collected and processed with respect to different values of at least one appropriate diversity parameter. For example, channel responses may be combined over multiple time and/or frequency intervals.

Interference rejection can be accomplished using subspace orthogonalization techniques. Interference rejection can be facilitated when the number of spatial subspaces used for communication is less than the total available number of spatial subspaces. For example, weak spatial directions may not be utilized by the communication system. Unused spatial subspaces may also result from unequal numbers of transmitters and receivers. Furthermore, non-spatial subspaces may also be used to mitigate interference. In any of these cases, one or both ends of the communication link have extra degrees of freedom that can be used to mitigate interference. In the case where interfering signals are canceled between different directional networks, the directional networks can provide additional degrees of freedom. For example, nulls in the beam patterns generated by the networks can be positioned in directions of undesired interferers. Other interferers that are not nulled in each of the beam patterns are canceled when the network outputs are weighted and combined.

A well-known method of mitigating interference is to "whiten" the effect of interference. For example, interference can be minimized and spread evenly across diversity parameters and/or diversity-parameter subspaces. An interleaving process typically distributes symbols across one or more diversity parameters. Interleaving helps

distribute the effects of channel estimation errors, distortion, and interference localized to a particular diversity-parameter value or set of values so that decoder error is reduced.

In one embodiment, an interference covariance matrix is averaged across all bins. The covariance matrix may also be averaged across a predetermined time interval to provide an average spatial interference covariance matrix that is independent of any particular bin. In some cases, it may be advantageous to add a scaled matrix identity term to estimates of the interference covariance matrix for reducing the sensitivity of interference-mitigation processes to covariance estimation errors.

Similar interference-mitigation techniques may be employed at a transmitter to reduce interference radiated to unintentional receivers. One possible interference-reduction technique is to whiten transmitted interference across diversity parameters and/or diversity-parameter subspaces.

The maximum number of diversity-parameter subchannels is related to either or both the number of transmitted diversity-parameter value sets and the number of received diversity-parameter value sets. Deep fades and interference can render some subchannels unusable. In some cases, transmitter weights and/or receiver weights may be optimized to select one or more subchannels that maximize average received power for a predetermined set of communication channels (i.e., connections between transmitters and receivers). Another criteria for selecting subchannels is to maximize received power subject to constraints on the average interference power radiated to unintentional receivers and/or received from unintentional transmitters.

17. System Implementation and Measured Results

FIG. 45 shows a behavioral model of a field-programmable gate array (FPGA) chip that was constructed for a CI transceiver. The FPGA provides a sub-physical layer core for a multi-mode transceiver. This particular design enables the transceiver to switch between DS-CDMA, TDMA, OFDM, and MC-CDMA. Furthermore, each of the transmission protocols has the superior capacity and signal-quality benefits provided by the CI signal architecture.

CI pulses are constructed from a set of waves stored in a wave look-up table. Thus, this implementation avoids the use of Fourier transform operations. CIMA pulse trains are constructed by combining time-offset CIMA pulses, which are stored in RAM.

The pulses are provided with complex weights to construct desired time-domain characteristics. These weights shape the physical-layer protocol and may help shape the MAC-layer protocol. In particular, weights may be used to shape signals to conform to various standards, such as Blue Tooth. Weights may also include information symbols. The weights may be used to construct preambles, headers, IP addresses, synchronization sequences, control symbols and the like.

The carriers that comprise the pulses may be modulated with complex symbols to convey information, construct physical-layer protocols, shape the pulses, etc. Furthermore, the carriers may be selected, adjusted, weighted, or otherwise adapted to effect addressing, power control, channel selection, multiple access, synchronization and the like. Thus, it should be appreciated that physical-layer (and sub-physical layer) attributes may be used to effect higher-layer (e.g., MAC layer) processes.

The FPGA includes two CI-signal generators. One signal generator provides a CI pulse train that is modulated with weights and transmitted. The second signal generator provides a reference signal that is combined with a received signal in a correlator receiver. The FPGA is capable of generating carriers within a 100-MHz baseband. Up to 512 carriers may be selected and approximately 80 simultaneous channels can be transmitted and/or processed in the receiver.

FIG. 46A illustrates a time-domain CI pulse generated by the FPGA. The pulse is formed by a superposition of ten unweighted carrier frequencies that are spaced about 100 kHz apart. A frequency-domain plot of the pulse is shown in FIG. 46B.

FIG. 47A illustrates a simplified DS-CDMA signal. This signal was formed using a 32-chip Hadamard-Walsh code with 32 carriers. The carriers are separated by approximately 120 kHz. A Fourier transform of the DS-CDMA signal (shown in FIG. 47B) illustrates how applying binary weights to CI pulses is equivalent to applying complex weights to the carrier signals. In fact, each unique time-domain signature has a corresponding unique set of carrier weights. Thus, separation of received encoded signals may be performed by correlating a received signal with a coded reference pulse train or correlating the received signal with each of a set of complex-weighted carriers.

FIG. 48 illustrates a frequency-domain plot of a transmitted 32-chip Hadamard-Walsh signal in a multipath environment. Multipath causes the complex values of the

carriers to change. Thus, multicarrier signals transmitted from different spatial locations are provided with unique "multipath codes." These multipath codes may be used to separate signals having similar frequency-domain and time-domain characteristics.

Fig. 49A shows a time-domain plot of a 64-chip Hadamard-Walsh code generated with 32 carriers. In this case, the CI pulses overlap to provide 64 quasi-orthogonal pulses. A frequency-domain plot of the CI pulses is shown in FIG. 49B.

18. Scope of the Invention

In the preferred embodiments, several kinds of carrier interferometry and spatial processing are demonstrated to provide a basic understanding of diversity reception and spatial demultiplexing of signals having an underlying CI architecture. With respect to this understanding, many aspects of this invention may vary. For example, signal spaces and diversity parameters may include redundantly modulated signal spaces. Descriptions of spatial processing may be applied to processing methods for non-spatial diversity parameters. Descriptions of systems and methods using spatial subspaces may be extended to systems and methods that use non-spatial subspaces.

For illustrative purposes, flowcharts and signal diagrams represent the operation of the invention. It should be understood, however, that the use of flowcharts and diagrams is for illustrative purposes only, and is not limiting. For example, the invention is not limited to the operational embodiment(s) represented by the flowcharts. The invention is not limited to specific signal architectures shown in the drawings. Instead, alternative operational embodiments and signal architectures will be apparent to persons skilled in the relevant art(s) based on the discussion contained herein. Also, the use of flowcharts and diagrams should not be interpreted as limiting the invention to discrete or digital operation. In practice, as will be appreciated by persons skilled in the relevant art(s) based on the herein discussion, the invention can be achieved via discrete or continuous operation, or a combination thereof. Further, the flow of control represented by the flowcharts is provided for illustrative purposes only. As will be appreciated by persons skilled in the relevant art(s), other operational control flows are within the scope and spirit of the present invention.

Exemplary structural embodiments for implementing the methods of the invention are also described. It should be understood that the invention is not limited to the

particular embodiments described herein. Alternate embodiments (equivalents, extensions, variations, deviations, combinations, etc.) of the methods and structural embodiments of the invention and the related art will be apparent to persons skilled in the relevant arts based on the teachings contained herein. The invention is intended and adapted to include such alternate embodiments. Such equivalents, extensions, variations, deviations, combinations, etc., are within the scope and spirit of the present invention.

Signal processing with respect to sinusoidal oscillating signals are described herein. Those skilled in the art will recognize there are other types of periodic oscillating signals that could be alternatively used, including, but not limited to sinusoids, square waves, triangle waves, and arbitrary waveforms.

The foregoing discussion and the claims that follow describe the preferred embodiments of the present invention. With respect to the claims, it should be understood that changes could be made without departing from the essence of the invention. To the extent such changes embody the essence of the present invention, each naturally falls within the breadth of protection encompassed by this patent. This is particularly true for the present invention because its basic concepts and understandings are fundamental in nature and can be broadly applied.